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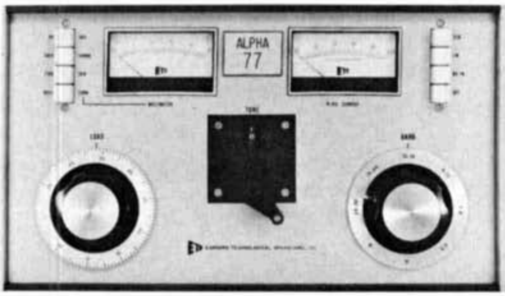
DECEMBER 1973

this month

HOW RUGGED IS AN ALPHA 77?



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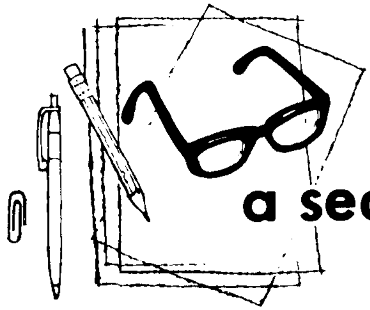

a second look

by jim fisk

November 27th *marks the fiftieth anniversary* of one of amateur radio's most memorable events — the first two-way amateur communications across the Atlantic Ocean. It was a hard-won goal, its path marked with failure and frustration, *but when the Atlantic, at last, had been* spanned, it was conquered by *short-wave* amateur radio, on wavelengths that previously were considered to be useless.

The first Transatlantic tests, in December, 1920 were a dismal failure, as were a second series of tests conducted in February, 1921. The 250 or so British stations which were listening for prearranged signals from the United States on a wavelength of 200 meters jammed each other so badly with radiations from their own regenerative receivers that they couldn't hear any signals from across the pond!

A third Transatlantic test was scheduled for December, 1921. In November, Paul Godley, 2XE, designer of the famous Paragon receiver, sailed from New York with two receivers under his arm — one a standard variometer regenerative set with two stages of audio amplification, the other a 10-tube superheterodyne built especially for the tests. With this superhet and a Beverage antenna installed on the bleak Androssan moor on the coast of Scotland, Godley heard the first stateside signals coming through in the wee hours of the morning on December 8th.

A year later, two European stations, F8AB in Nice, and G5WS in London, were heard along the east coast of the United States, but two-way communications were as elusive as ever.

A fourth series of Transatlantic tests were scheduled for late 1923. However, these carefully laid plans were totally upset by the enterprise of one man, Leon Deloy, F8AB. Deloy came to the states during the summer of 1923 where he met *with John Reinartz, 1XAM, and Fred* Schnell, 1MO. Deloy picked up a lot of valuable advice from his talks with Reinarts and Schnell, and before returning to France he acquired a new Grebe receiver and the details of a "trick" *circuit which, he was told, would "go* down to about 100 meters." Up until that time all the Transatlantic tests had been conducted on a wavelength of 200 meters.

Deloy put his new 100-meter station *on the air in early autumn, and having* satisfied himself that everything was in working order, cabled Schnell that he would transmit on 100 meters between 0200 and 0300 GMT on November 26, 1923. The signals from F8AB were heard by Schnell and Reinartz almost from the first dot he transmitted, but the Americans were not ready to transmit back. Unlike Deloy, who presumably did not think it was necessary to obtain official permission to transmit on such a short wavelength, Schnell had to seek the necessary authority from the Radio Supervisor in Boston.

On November 27th Schnell received special permits from Boston for himself and Reinartz. Late that night (early morning in Europe) they were both on the air. For an hour Deloy called the United States and then sent two messages. At 0330 GMT he signed off and asked for acknowledgement. Long calls followed from 1MO and 1XAM. Then came the eagerly awaited reply — Deloy had heard both stations clearly. Reinartz was asked to stand by as Deloy transmitted to Schnell, "R R QRK UR SIGS QSA VERY ONE FOOT FROM PHONES ON GREBE FB OM HEARTY CONGRATULATIONS THIS IS A FINE DAY — PSE QSL. It was, indeed, a fine day.

Jim Fisk, W1DTY
editor

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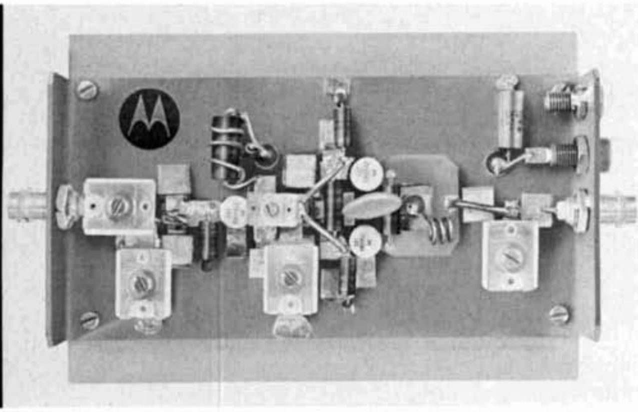


a solid 80 watts for two meters

Construction details for high-performance, solid-state two-meter power amplifiers

John Hatchett, Motorola Semiconductor Products, Inc.

Evaluation results, component layout and construction information for two 80-watt vhf power amplifiers are described. These solid-state amplifiers can be used to boost two-meter output power levels to 80-watts. Both units have been designed to operate from a dc supply voltage of 12.5 volts with 50-ohm source and load impedances. The 12.5-volt power requirements are easily adapted to fixed or mobile operation.

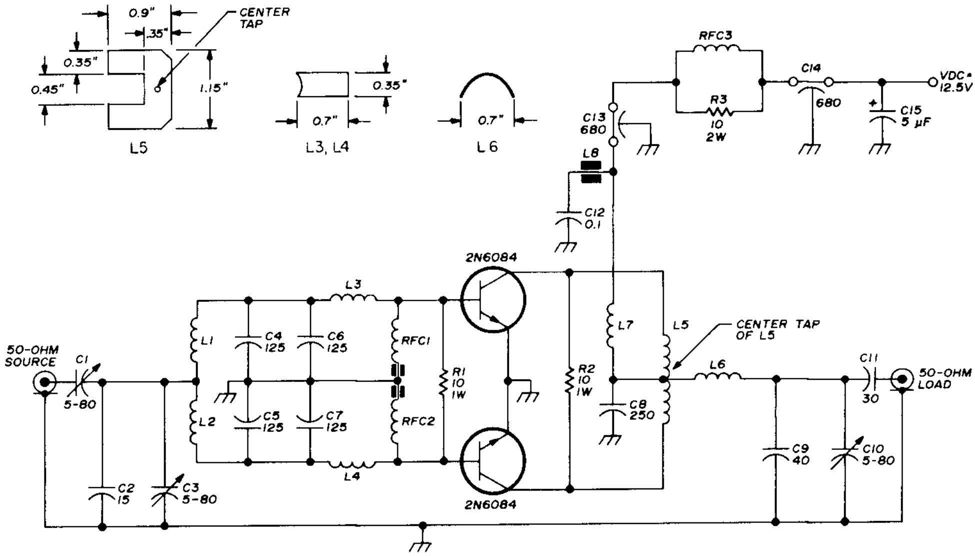
One of the amplifiers is a single-stage design using two 2N6084 transistors combined with simple LC components (**fig. 1**). It can be tuned to operate from 144 to 175 MHz, and requires a typical input power level of 20 watts for 80-watts output at 144 MHz.

The second amplifier uses the same output stage design, but adds a 2N6083 transistor driver stage (**fig. 2**) to reduce input drive requirements. This design is also tunable from 144 to 175 MHz, and will provide 80-watts output power at 144 MHz with only 2.5-watts of drive.

Six single- and two-stage amplifiers have been constructed and evaluated with similar performance exhibited by the amplifiers in each group. Typical values for the more important amplifier characteristics are shown in **table 1** and in **figs. 3, 4, 5** and **6.** The amplifiers have also been subjected to momentary open- and short-circuit load conditions without damage to the transistors.

design philosophy

The amplifiers have been designed to be efficient, reliable and stable without sacrificing simplicity. All amplifier stages are of the common-emitter configuration, operated class-C. Two 40-watt 2N6084 transistors have been used in the high-power output stage to provide excellent heat distribution at full power. Combining the two 2N6084 devices is accomplished with LC signal splitting/combining techniques. For the single-stage amplifier the combinations of L1 and L3 and L2 and L4 split the signal, and inductance L5 recombines the signals. The two-stage amplifier uses L4 and L5 for signal splitting and L6 for combining. These inductors provide impe-



C1,C3,C10	5-80 pF trimmers (ARCO 462)
C2	15 pF metal clad (Underwood Electric type J-101*)
C4,C5,C6,C7	125 pF metal clad (Underwood Electric type J-101)
C8	250 pF metal clad (Underwood Electric type J-101)
C9	40 pF metal clad (Underwood Electric type J-101)
C11	30 pF metal clad (Underwood Electric type J-101)
C12	0.1 µF, 75 V ceramic disc
C13,C14	680 pF feedthrough (Allen Bradley type FA5C)
C15	5.0 µF, 25v, aluminum electrolytic
L1,L2	2½ turns number-16, 0.2" ID (60 nH)
L3,L4	cut from 0.031" single-sided G10 circuit board (5 nH)
L5	cut from 0.031" single-sided G10 circuit board (8 nH to center tap)
L6	number-12 wire, approximately 1.1" long (10 nH)
L7	3 turns number 14, 0.25" ID (50 nH)
L8	ferrite bead (Ferroxcube 5659065/3B)
RFC1,RFC2	0.15 µH molded choke with Ferroxcube 5659065/3B ferrite bead on ground lead
RFC3	10 turns number-14 wire wound around R3

fig. 1. Schematic diagram of the single-stage, 80-watt, 144-MHz power amplifier. Circuit is built on 0.062" single-sided G10 circuit board. Performance of this amplifier is graphed in figs. 3 and 5.

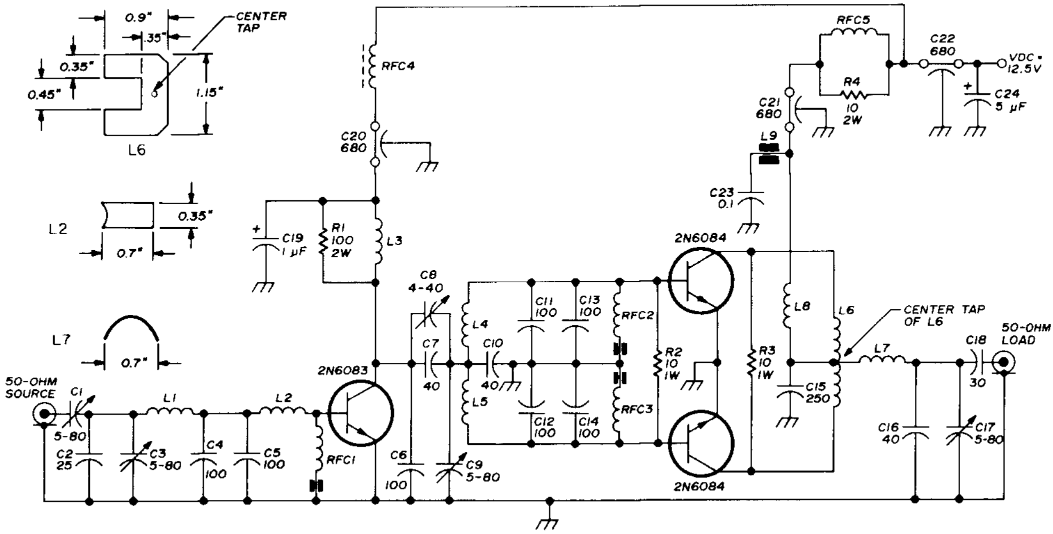
dance transformation, isolation between devices and minimize unequal load sharing.

Low-loaded Q impedance matching network designs have been used to maximize bandwidth and to minimize insertion loss. This also results in reducing reflected voltage levels that can occur during high-output vswr conditions. A low-pass, low-loss, LC output filter can be used to provide additional attenuation of the harmonic components.

The two transistor types used in the amplifiers are part of the Motorola vhf land mobile series designed for 12.5-volt fm operation. They are multiple balanced-emitter transistors manufactured using the †Isothermal process technology to minimize temperature variations across the transistor chips. This process provides increased transistor protection over wide thermal and load vswr excursions. The devices are packaged in a 0.380 inch diameter, stripline-opposed-emitter stud package (case 145A-01).

*Underwood Electric & Mfg. Co., Inc., 148 South Eighth Avenue, Maywood, Illinois 60153.

†Trademark of Motorola Inc.



C1,C3, C9,C17	5-80 pF trimmers (ARCO 462)	L3	3 turns number-16 wound around R1 (60 nH)
C2	25 pF metal clad (Underwood Electric type J-101)	L4,L5	1.1" long number-14 wire, formed around 0.6" diameter cylinder (12 nH)
C4,C5,C6, C12,C13,C14	100 pF metal clad (Underwood Electric type J-101)	L2	cut from 0.031" single-sided G10 circuit board (5 nH)
C7,C10,C16	40 pF metal clad (Underwood Electric type J-101)	L6	cut from 0.031" single-sided G10 circuit board (8 nH to center tap)
C8	4-40 pF trimmer (Arco 403)		
C15	250 pF metal clad (Underwood Electric type J-101)	L7	number-12 wire, approximately 1.1" long (10 nH)
C18	30 pF metal clad (Underwood Electric type J-101)	L8	3 turns number-14, 0.25" ID (50 nH)
C19	1.0 μF tantalum	L9	ferrite bead (Ferroxcube 5659065/3B)
C20,C21,C22	680-pF feedthrough (Allen Bradley type FA5C)	RFC1, RFC2, RFC3	0.15 μH molded choke with Ferroxcube 5659065/3B ferrite bead on ground lead
C23	0.1 μF, 75 V ceramic disc		
C24	5.0 μF, 25V, aluminum electrolytic	RFC4	ferrite choke (Ferroxcube VK200 19/4B)
L1	1 turn number-16, 0.25" ID (18 nH)	RFC5	10 turns number-14 wound around R4

fig. 2. Schematic diagram of the two-stage, 80-watt, 144-MHz power amplifier. Circuit is built on 0.062" single-sided G10 circuit board as shown in the photograph. Performance of this amplifier is shown in figs. 4 and 6.

To achieve the 80-watt power level, it is imperative that low-loss matching network components be used. It is also necessary that these components be characterized for the desired operating frequencies. Suitable low-loss coils can be made with a small length of wire, ribbon conductor or printed circuit board material. Economical capacitors for efficient high-power operation at 2 meters are more difficult to obtain. All fixed capacitors in the amplifiers, 250 pF or less in value, are Underwood mica dielectric units. The effective capacitance of these components at 2-meters will deviate only slightly from the low frequency value for nominal capacitance values up to approximately 60 pF. Larger capacitors of this type are characterized for operation at the selected frequency.

construction

A full description of all necessary components for building the amplifiers along with the schematic diagrams are shown in **figs. 1** and **2**. Care must also be given to the physical location of the components. The photograph and the scale drawings in **figs. 7** and **8** can be used to determine proper component placement. For the sake of simplicity, only those components necessary to establish the basic amplifier layout have been included in the drawings.

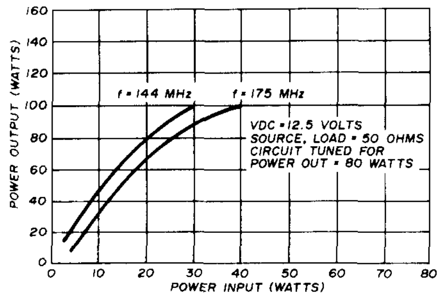


fig. 3. Power output vs power input for the single-stage amplifier.

table 1. Amplifier performance for a dc supply voltage of 12.5 volts.

		single-stage design	two-stage design
Power output (watts)		80	80
Power input (watts)	144 MHz	20	2.5
	148 MHz	21	2.6
	165 MHz	23	3.5
	175 MHz	26	5.5
Power gain at 144 MHz	(dB)	6.0	15.1
Dc current (amperes)	output stage	8.5	8.5
	driver stage	—	2.5
Harmonic attenuation	(dB)	20	20
Stability	Amplifiers are stable for input drive levels from zero to 30% overdrive and for supply voltages from 8.0 to 15.5 volts dc.		
Ruggedness	With 80 watts power output into 50 ohms, no transistor damage from open- and short-circuit load conditions for all phase angles		

The amplifiers are built on 0.062 inch, single-sided, G10 circuit board with the components mounted on the ground plane side. In each case, the ground plane is continuous except for interruptions for the transistor and feedthrough capacitor (C13, C14 and C20, C21, C22) mounting holes. Coils L3, L4 and L5 of the single-stage amplifier isolate the transistor base and collector contacts from the ground plane. Coils L2 and L6 accomplish this function in the two-stage design. In addition, four small pads of 0.31-inch, G10 circuit board are used to provide isolation for the 2N6083 collector, the base of each 2N6084 and capacitor C7.

To prevent physical damage to the transistor stud package, the following precautions should be observed:

A. The maximum torque ratings for the mounting nut must not be exceeded (6.5 inch-pounds for the 2N6083 and 2N6084 devices).

B. The nut should be placed on the stud and tightened to the specified torque before soldering the transistor leads to the circuit. After the nut is properly torqued, a slightly downward pressure can be exerted on the leads to place them in contact with the circuit board connection points. The objective is to prevent an upward force being applied to the leads near the case body.

thermal considerations

The amplifiers must be provided with

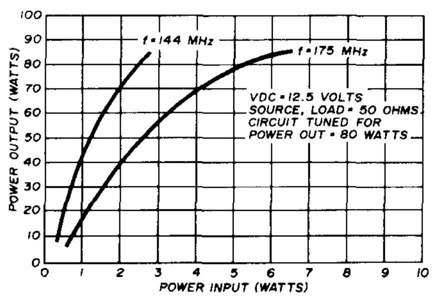


fig. 4. Power output vs power input for the two-stage amplifier.

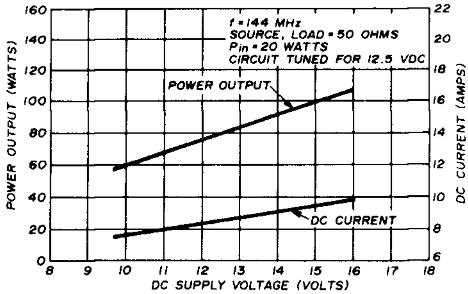


fig. 5. Amplifier power output and dc current vs dc supply voltage for the single-stage amplifier.

heat sinks capable of keeping the transistor junction temperatures below their specified maximum temperature of 200°C. This requires extremely good thermal design and construction practice. A smooth heat-sink surface is required to maximize heat-sink to transistor case contact area. A proper amount of thermal-joint compound must be used between heatsink and transistor case interface to improve thermal transfer to the heatsink. Wakefield type 120, Thermoalloy *Thermacote,* Dow Corning type 340 or other thermal compounds exhibiting similar low thermal resistance properties are recommended. The heatsink must have a thermal resistance low enough to adequately transfer the heat from the transistor case to surrounding air.

Limiting the transistor junction temperatures to a maximum of 180°C during continuous operation into a 50-ohm load requires a heat-sink thermal resistance specification of less than 1.7°C/watt for the output stage devices at 60°C ambient. For an ambient of 30°C, the heat-sink thermal resistance requirement can be relaxed to approximately 2.3°C/watt. Similar operating conditions require the 2N6083 driver transistor heat-sink thermal resistance to be less than approximately 6° and 8°C/watt for ambient temperatures of 60° and 30°C, respectively.

Duty cycle operation, such as 1-minute on/3-minutes off, will significantly reduce the heat-sinking requirements. If operation into mismatched loads is anticipated, the heat-sink thermal resistance values must be reduced to account for the radical increase in transistor power dissipation that can occur with these operating conditions.

Several economical aluminum heatsinks are available with thermal resistance values in the order of 3°C/watt. These would be adequate for use with the amplifiers in most applications, since a 50-ohm load is used and continuous operation capability is not required. More expensive heatsinks can provide thermal resistance values less than 1°C/watt. **Table 2** provides a brief description for some of the commercially available units.

amplifier adjustment

An amplifier alignment test set-up is shown in **fig. 9.** Initial amplifier tuning should be started with reduced supply voltage (approximately 8 volts) and reduced drive levels to prevent excessive device dissipation. For 144-MHz operation, a reasonably good starting point would be to set all variable capacitors approximately ½-turn from the fully closed (maximum capacity) position. During alignment, you may carefully touch each transistor case to detect excessive power dissipation in any of the transistors. Each transistor case should feel warm, but not too hot.

If a spectrum analyzer is available, it should be used to monitor the output signal during tuneup to verify proper alignment and to indicate the presence of low-frequency oscillations that can occur if the amplifiers are significantly mal-

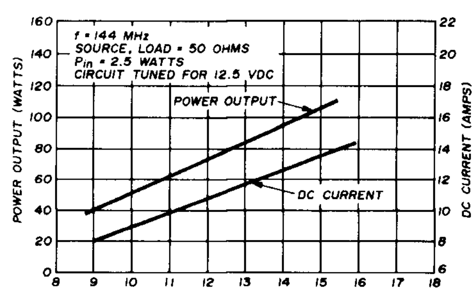


fig. 6. Amplifier power output and dc current vs dc supply voltage for the two-stage amplifier.

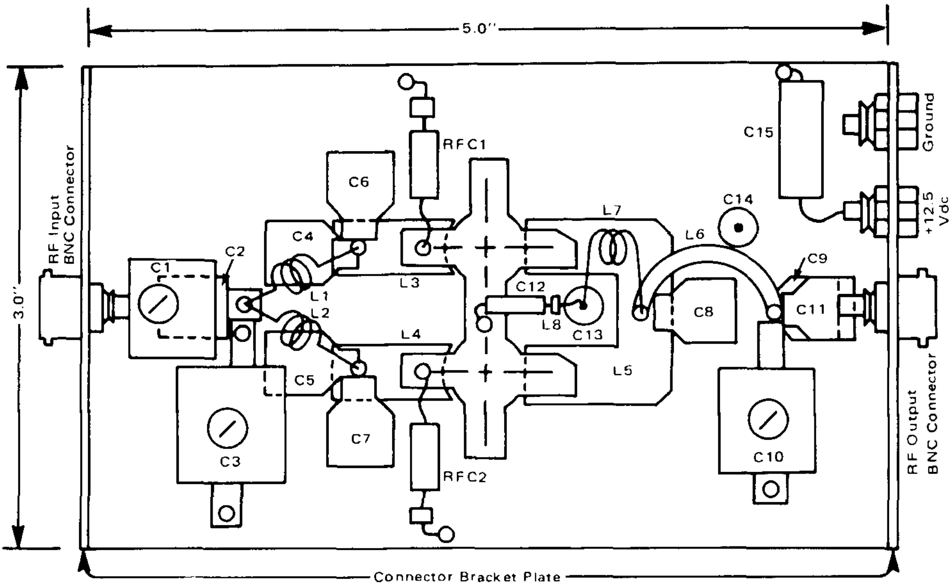


fig. 7. Component location for the single-stage amplifier. Some components have been omitted for clarity. Transistor mounting holes are 0.80", center to center. RFC3 and R3 are mounted on rear side of board.

adjusted. An oscilloscope connected to the dc voltage line (for example, at the top of the 5-μF filter capacitor, C15 or C24) can also be used to provide useful information on the presence of low-frequency oscillations. The scope probe will usually pick up enough of the two-meter signal energy to provide a signal display on the CRT.

If low-frequency oscillations are not present, the two-meter signal display will be constant in amplitude. If a low-frequency oscillation (typically less than 10 MHz) is present, it will show up as amplitude variations on the two-meter display. The frequency of the amplitude variations correspond to the frequency of the oscillation. High-frequency oscilloscopes (100 MHz) will provide a good display of the two-meter signal. Low-frequency oscilloscopes (20 MHz) are not capable of showing the two-meter signal itself, but they can be useful in determining if a low-frequency amplitude variation (envelope) is present on the two-meter signal. Any oscillation should be eliminated by adjusting the amplifier variable capacitors.

Single-stage amplifier. Start with low drive level (approximately 2 to 5 watts)

table 2. Summary of commercial heatsinks suitable for use with the 80-watt, two-meter amplifier. Thermal resistance values (column 2) are for natural convection except for the Wakefield FC-502 and FC-503 units, which are for 10 cubic feet/minute air flow.

part number	approximate thermal resistance (°C/watt)	description
WEI Corp. 3110	2.5	aluminum, 1.3" x 4.0" x 1.5" & 3.0"
WEI Corp. 3164	2.5	aluminum, 1.0" x 4.12" x 1.5" & 3.0"
Thermalloy 6169	2.5	aluminum, 1.3" x 4.12" x 3.0"
Wakefield NC-641	2.5	aluminum, 1.0" x 4.12" x 3.0"
Wakefield FC-502	0.45	copper, 1.75" x 3.5" x 1.75"
Wakefield FC-503	0.35	copper, 1.75" x 3.5" x 3.5"

WEI Corporation, P. O. Box 10577, Santa Ana, California 92705
Thermalloy Inc., 8717 Diplomacy Row, Dallas, Texas 75247
Wakefield Engineering Inc., Wakefield, Massachusetts 01881

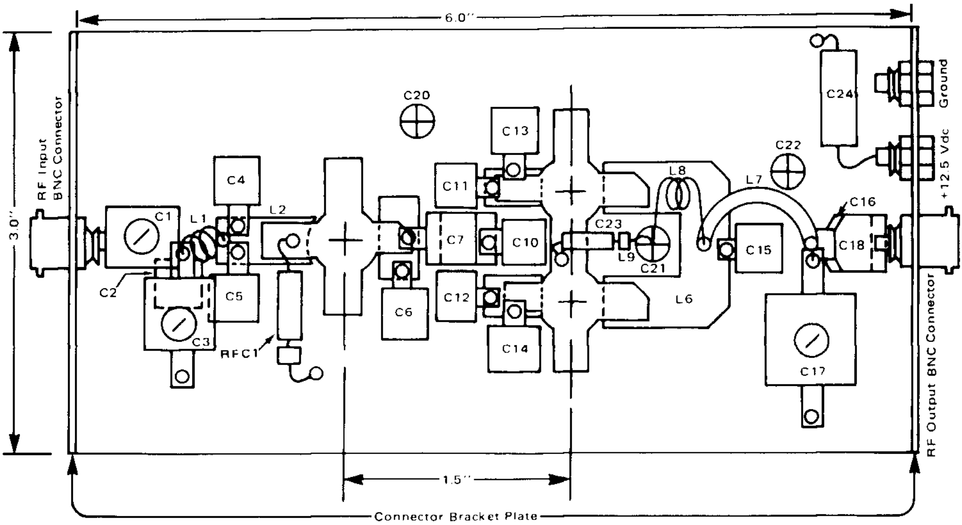


fig. 8. Component location for the two-stage amplifier. Some components have been omitted for clarity. Transistor mounting holes are 0.80″, center to center. RFC4, RFC5 and R4 are mounted on back of board.

large enough to turn on the transistors (indicated by the flow of dc collector current). Then adjust capacitor C10 for maximum output and C1 and C3 for minimum reflected power to the drive source as indicated by the swr bridge. Increase the supply voltage to 12.5 volts after this initial adjustment, and continue to increase the input drive power to approximately 8 to 12 watts while adjusting C10 first and then C1 and C3 as before.

Increase the drive power to approximately 20 watts, and tune for rated power output in a similar manner. After tuning for rated output power, capacitor C10 can be increased slightly in capacitance. This will minimize the required dc current with only a slight degradation, approximately 0.1 dB or less, in power output.

Two-stage amplifier. Start with low drive (approximately 0.25 to 0.5 watt) large enough to turn on the 2N6083 stage as indicated by the flow of dc collector current. Then adjust C8, C9 and C17 for maximum output power and C1 and C3 for minimum reflected power to the drive source as indicated by the swr bridge. Increase the supply voltage to 12.5 volts after this initial adjustment, and increase the input drive power to approximately 1.0 to 1.5 watts while adjusting C17 first and then C8, C9 and C1, C3 as before.

Now, increase the drive power to approximately 2.5 watts and tune for rated power output. After tuning for rated output power, capacitor C17 can be increased slightly in capacitance. This will minimize the output stage dc current requirement with only a slight degradation, approximately 0.1 dB or less, in power output.

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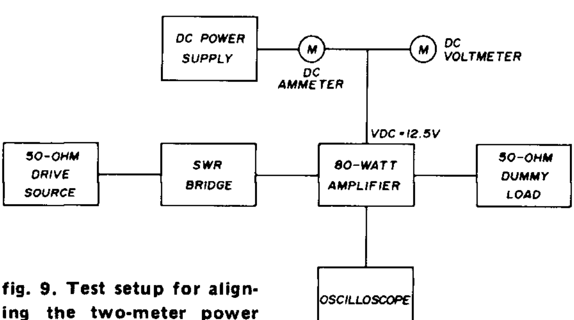


fig. 9. Test setup for aligning the two-meter power amplifiers. Spectrum analyzer and power attenuators should be used if available.

More Details? CHECK—OFF Page 126

crystal controlled AFSK generator

Complete construction details for the RY-170 — an AFSK synthesizer for 170-Hz shift

Howard Nurse, W6LLO, Palo Alto, California 94306

How would you like to generate precise RTTY audio tones without the need for a counter to establish the correct frequencies? Many years ago, when faced with the same problem, I recall an attempt to use a guitar to help adjust an AFSK oscillator! With that technique leaving something to be desired, I often thought how nice it would be to have an oscillator which could generate the correct frequencies without adjustment, while not breaking the bank in the process. Enter the RY-170, described here.

It wasn't until recently that surplus integrated circuits have made inexpensive frequency synthesis techniques possible. Start with a surplus crystal, divide by the correct ratios to generate 2125 and 2295 Hz, add a simple active bandpass filter, and for less than ten dollars you can have a 170-Hz shift synthesizer in your RTTY system.

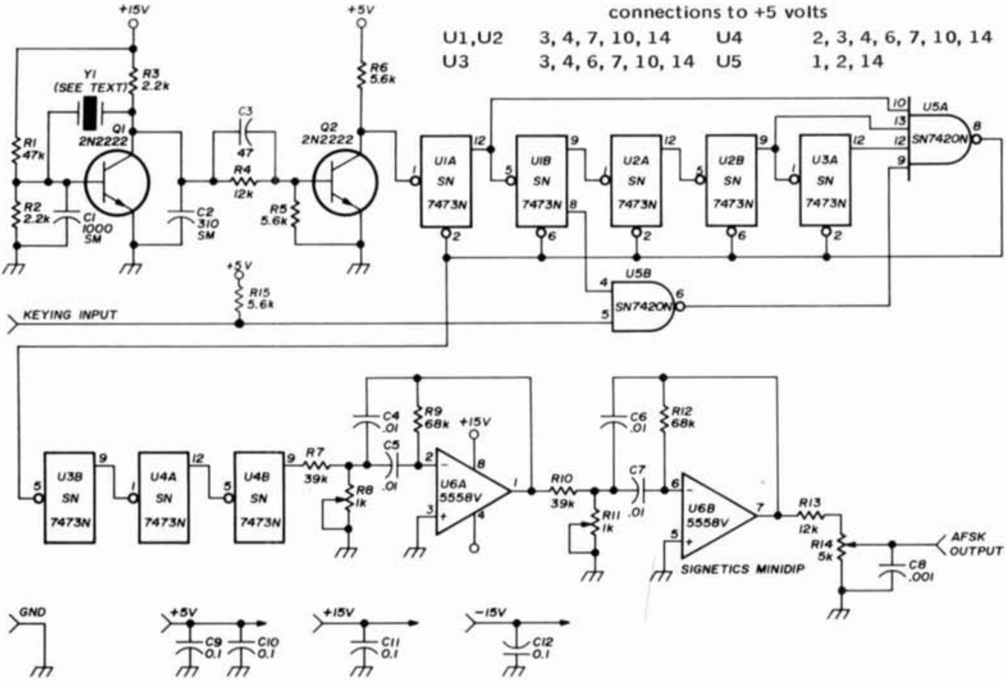


fig. 1. Schematic diagram of the RY-170 AFSK generator. Pin 11 of ICs U1, U2, U3 and U4, and pin 7 of U5, are connected to ground.

design

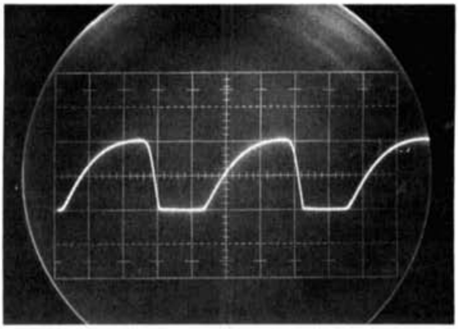
The following goals were established prior to starting the RY-170 project: Generation of 2125- and 2295-Hz tones from one crystal, low output distortion (THD), minimal keying overshoot, control from a TTL compatible input, use of inexpensive components, and easily duplicated printed circuit board. The resulting design is shown in **fig. 1**, while the photographs show various views of the completed unit. A summary of the RY-170 specifications is given in **table 1**.

circuit description

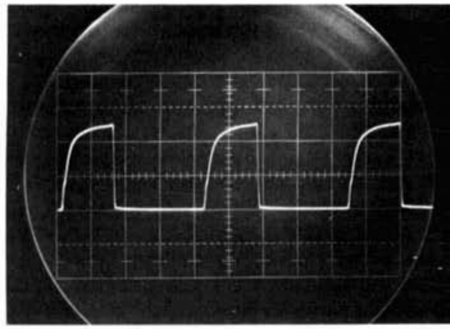
The heart of the AFSK synthesizer is an oscillator using a surplus FT-241 crystal and transistor Q1 in a modified Pierce circuit. A channel 48 (459.259 kHz) crystal will yield output frequencies accurate to approximately 2 Hz, while preserving the relative shift (170 Hz) to within 0.1 Hz. If you desire even greater accuracy (with a slight increase in cost), an FT-241 crystal can be ordered which has been adjusted to the correct fre-

Rear view of the RY-170 AFSK generator with the top cover removed.

fig. 2. Waveforms of the RY-170 AFSK generator.



A. Collector Q1
horizontal scale, 0.5 microseconds/cm vertical scale, 5.0 volts/cm ground 2 cm from bottom

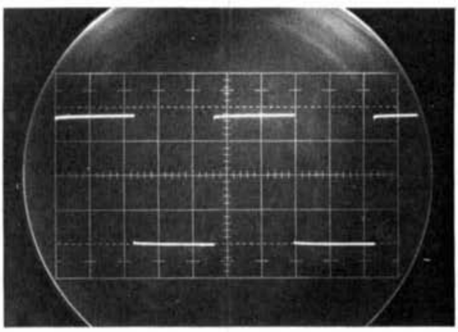


B. Collector Q2
horizontal scale, 0.5 microseconds/cm vertical scale, 2.0 volts/cm ground 2 cm from bottom

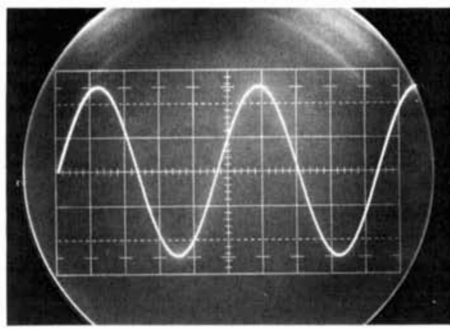
quency of 459.000 kHz.* The output waveforms from the oscillator and its buffer are shown in **figs. 2A** and **2B**, respectively.

Transistor Q2 interfaces the output of the oscillator to the divider input. JK flip-flops U1 through U3A and NAND gate U5 are wired as a programmable divider whose divide ratio depends on the logic state of the synthesizer input. When the input is grounded, the divide ratio is 25 (2295 Hz); when the input is high, the ratio is 27 (2125 Hz). The programmable portion of the divider is followed by an additional divide-by-eight circuit consisting of flip-flops U3B and U4. The output waveform of the complete divider is a TTL square wave, shown in **fig. 2C.**

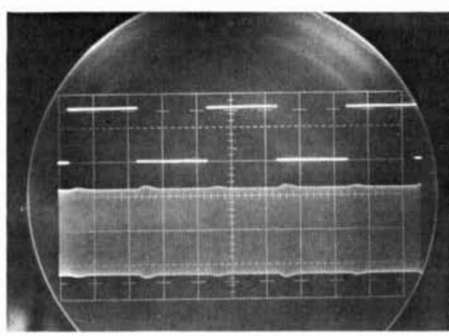
The output bandpass filter, necessary to extract the fundamental frequency



C. Divider output
horizontal scale, 100 microseconds/cm vertical scale, 1.0 volt/cm ground 1 cm from bottom



D. Filter output
horizontal scale, 100 microseconds/cm vertical scale, 0.5 volts/cm ground at center



E. Top trace — TTL-compatible keying waveform horizontal scale, 5 milliseconds/cm vertical scale, 2 volts/cm
Bottom trace — RY-170 output horizontal scale, 5 milliseconds/cm vertical scale, 0.2 volts/cm

*JAN Crystals, 2400 Crystal Drive, Ft. Myers, Florida 33901.

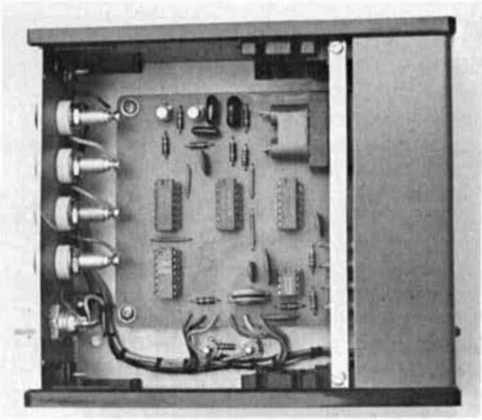
from the divider output, consists of dual-operational amplifier U6 and its associated components. The filter was designed with a Q of 10; raising the Q would result in increased filter sensitivity to component tolerances and increased keying overshoot, while lowering the Q would raise the THD.

adjustments

Resistors R8 and R11 are used to adjust the filter center frequency to pass the two tones. Resistors R8 and R11 should be adjusted, one for each tone, so that the transmitter has equal output power for 2125- and 2295-Hz inputs. The sinusoidal output waveform of the bandpass filter is shown in **fig. 2D** while the keying characteristics are shown in **fig. 2E.**

The output amplitude of the synthesizer can be adjusted with resistor R14 to match the transmitter audio requirements. R14 should be adjusted in conjunction with R8 and R11 to ensure that the audio stages of the transmitter are not overloaded.

Because the RY-170 is part of a larger system, I decided to use a common power supply for all accessories. If it is desired to use an internal supply with the AFSK board, a regulated supply meeting the requirements of the circuit can be used (see specifications in **table 1**).



Component layout of the RY-170 AFSK generator. A printed-circuit board is available from the author.

table 1. Specifications for the RY-170 170-Hz AFSK synthesizer.

Absolute frequency	459.000-kHz crystal: Mark: 2125 ±1 Hz Space: 2295 ±1 Hz
	Channel-48 crystal: Mark: 2126 ±1 Hz Space: 2296 ±1 Hz
Frequency shift	170.0 ±0.1 Hz
Output amplitude	0-1 volt p-p
Output distortion (THD)	less than 0.5%
Output keying overshoot	less than 5%
Power requirements	+5 ±0.5 volts at 100 mA +15 ±1 volts at 10 mA -15 ±1 volts at 10 mA
Keying	TTL high (open) for mark TTL low (ground) for space
Crystal frequency	459.000 kHz or Channel 48 FT-241
Divide ratio	Mark: 27 x 8 = 216 Space: 25 x 8 = 200

construction

The RY-170 AFSK synthesizer was constructed in a Ten-Tec JG-5 enclosure. The front panel was painted with Krylon 2021 (Oldsmobile green) which closely matches the color of the Heath SB-series. An LED is used as a pilot light, powered from the +5-volt power supply through a 220-ohm current-limiting resistor. The front and back panels are labelled with press-on letters and sprayed with Datak *Datakoat* for protection.

The circuit board is single-sided G-10 board and requires four jumpers.* I used Molex connector pins to hold the ICs although the ICs can also be soldered directly to the printed-circuit board. Dipped mica capacitors are required in the oscillator circuit, while high-stability capacitors (Orange Drop or polystyrene) should be used in the active filter.

There is something satisfying in knowing your shift is, and will remain, at 170 Hz. The RY-170 is one answer, and an economical one at that, to stable 170-Hz AFSK shift.

ham radio

*Drilled circuit boards and component layout information are available from the author for $5.50, postpaid.

wide range
rf signal generator

This wide range signal generator covers the range from 600 kHz to 12 MHz and features a built-in 1-kHz modulator

Hank Olson, W6GXN, P.O. Box 339, Menlo Park, California 94025

Most signal generators used by amateurs and other radio experimenters use the LC tuned oscillator in one form or another. Whether the exact circuit is a Colpitts or Hartley oscillator or some variation of these two basic designs, the output frequency is usually proportional to the inverse of the square root of the capacitance of the main tuning capacitor. That is, at least approximately

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Because of this relationship signal generator tuning is broken into bands, each of which encompasses a high- to low-frequency ratio of only 3 or 4 to one (check the dial on your own signal generator, and see). The 3 or 4 to one frequency ratio is a direct consequence of the fact that parallel-resonant LC circuits vary in frequency as $1/\sqrt{C}$, and most variable tuning capacitors have a maximum to minimum capacitance ratio of 20 to 1 or less.

The oscillator described here is *not* an LC type, and its output frequency is proportional to 1/C. Thus, the variable capacitor is capable of tuning the oscillator over a 20 to 1 range, from 600 kHz to 12 MHz! Since such a wide range of frequency is covered by a single 180° turn of the capacitor shaft, it is advisable to have a good sized dial for calibration. The largest dial of good quality that I could readily obtain (the biggest one in the junkbox) was a Millen 10035. The use of this dial is the only reason that the signal generator is as large as it is.

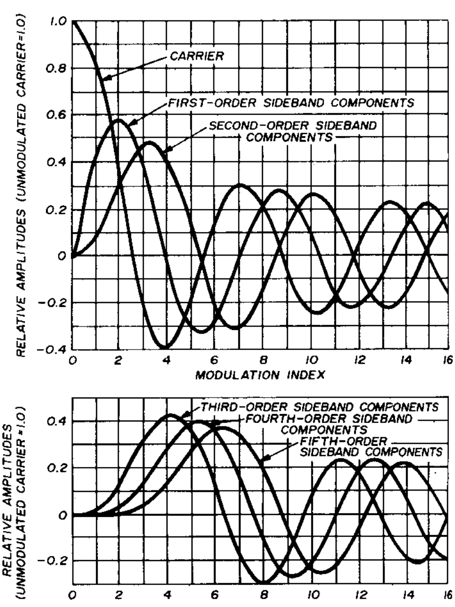


fig. 1. Graph of fm sideband components shows that the first-order sideband is nearly linear up to a modulation index of 0.5, and that higher-order sidebands are nearly non-existent at lower modulation indices.

circuit

A standard 365-pF broadcast tuning capacitor is used in the signal generator in conjunction with half of a relatively new Motorola IC, the Motorola MC4024P (or HEP 3805P). This IC is characterized as a dual voltage-controlled multivibrator, or vco. Since only half of the MC4024P is used to produce the rf output, the other half can be used to generate a 1-kHz modulation frequency.

The *waveforms produced* by both halves of the MC4024P are rectangular, and contain many harmonics. The harmonics of the rf oscillator section are not particularly troublesome, since *most* signal generators have appreciable harmonic content. However, it is desirable to filter the 1-kHz modulation waveform so that only one set of sidebands will be produced when the rf is modulated by the 1-kHz *signal. Actually, since small index frequency* modulation is used, there will be some small higher-order sidebands at 2 kHz and higher spacing around the rf carrier, but these will be insignificant if the generator is used within the bounds of narrowband frequency modulation (nbfm).

Nbfm has long since passed from the amateur scene, at least as a modulation method on the high-frequency bands. The main reason for nbfm's disfavor is that it is useful only for simulating amplitude modulation with 50% or smaller percentages. If higher indexes of fm are used, the higher-order sidebands rapidly increase and the signal no longer resembles a-m.

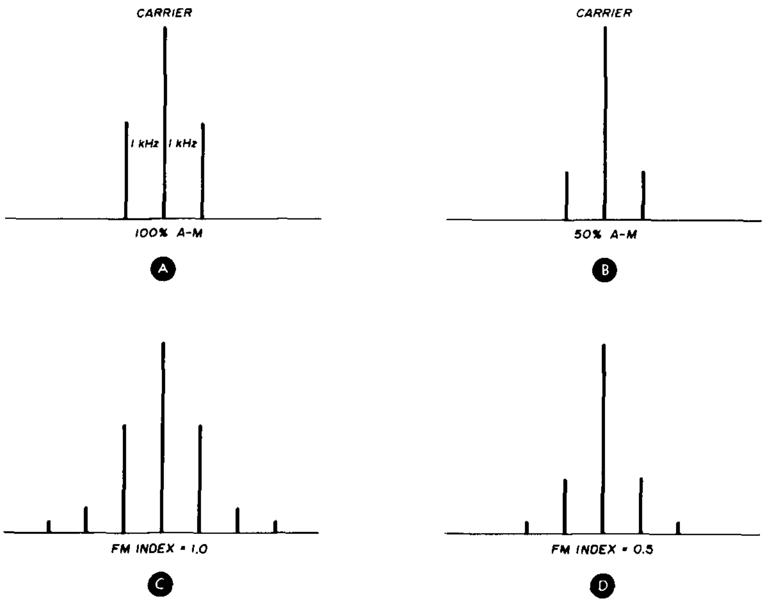


fig. 2. Frequency distribution for 50% and 100% amplitude modulation and fm signals with modulation indices of 0.5 and 1.0.

To see how this works, look at the graphs of **fig. 1.** Note that the graph representing the first-order sidebands is approximately *linear* up to a modulation index of 0.5 and that the higher-order sidebands are *almost* nonexistant at lower indices. **Fig. 2** shows the spectrum of a 100% a-m signal, a 50% a-m signal, a 1.0 order fm signal and a 0.5 order fm signal. *Note that the 1.0 order fm signal* succeeds in generating first-order sidebands comparable with those of the 100% a-m signal, but at the expense of producing 2nd and 3rd order sidebands of appreciable amplitude.

The 0.5 index fm signal gives a good approximation to a 50% a-m signal, with only small amplitude 2nd order sidebands. Since most amplitude modulated signal generators are only used at a-m percentages of about 50% (a standard measurement technique), you can use this 0.5 order fm signal to provide a simulated 50% a-m signal.

In fairness it must be mentioned that detection mode, where the receiver selectivity curve provides a frequency-to-amplitude conversion.

the circuit

The circuit of the signal generator is shown in **fig. 3.** Note that half of an MC3029P line-driver NAND gate follows each of the two multivibrators in the MC4024P. The line-driver NAND gates

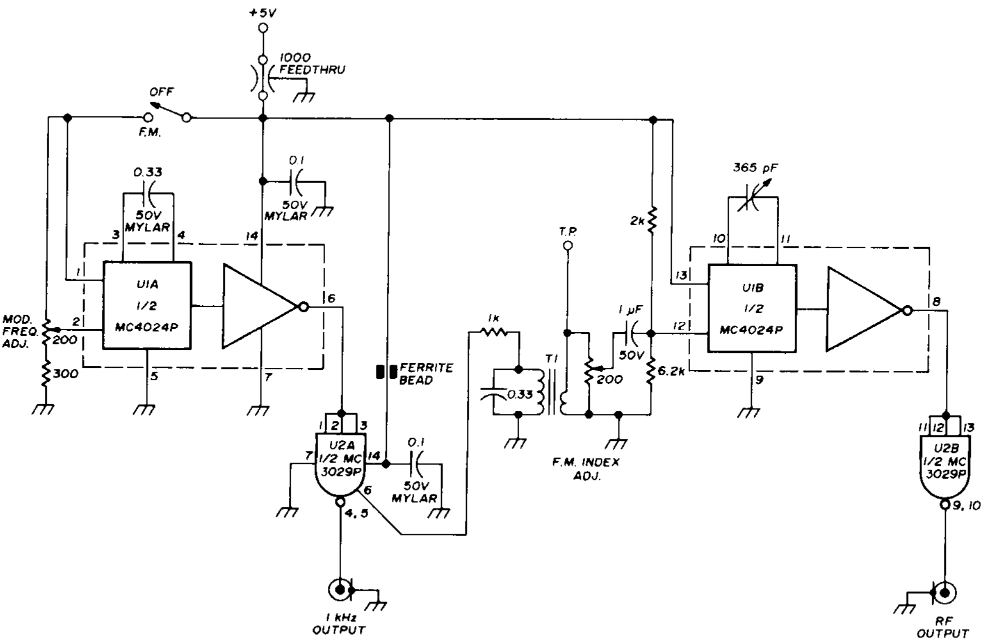


fig. 3. Circuit for the wide-range rf signal generator that covers from 600 kHz to 12 MHz. Integrated circuti U1A is the 1-kHz modulation oscillator; U1B is the rf oscillator. U2A and U2B are used as line drivers. Transformer T1 is an 88-mH toroid with a secondary consisting of 30 turns no. 28 enamelled wire wound over it.

nbfm approximates low percentage a-m *only* in the frequency domain. The signal is still fm since there is no variation in amplitude at the modulation rate. That this is true is immediately obvious because the entire system is made of digital ICs which are in essence *limiters*; that is, amplitude is constrained to be either 1 or zero.

Since the amplitude does *not* vary, a diode detector will, strictly speaking, be unresponsive to nbfm. However, most receiver systems having diode detectors will respond to nbfm by the slope-

provide isolation and the capability to drive 50-ohm lines with either 1-kHz or rf output. It must be remembered that the output of the generator is well over one volt peak-to-peak, even when terminated in 50 ohms, so an external attenuator is usually required.

The modulation frequency is determined by the parallel tuned circuit consisting of the 88-mH toroid, T1, and the 0.33μF capacitor across it. This is because the frequency of the 1-kHz oscillator is adjusted (by voltage control) to maximize the output at the test point. This occurs when the 1-kHz oscillator frequency matches the resonant frequency of the 88-mH-0.33-μF parallel tuned circuit. The voltage observed at TP with a scope should be about 0.2 volts p-p. If the resultant frequency at maximum TP voltage isn't close enough to 1-kHz to suit you, a somewhat different value of C (nominally 0.33-μF) will have to be used.

A simple but well-regulated power supply is shown in **fig. 4.** It uses a standard 6.3-Vac filament transformer and a full-wave bridge rectifier. The regulation is accomplished by one of the newer three-terminal IC voltage regulators of Fairchild, National or Motorola. The common terminal of each of these regulators is the case, so a good thermal connection to the chassis (for heat dissipation) is also the electrical ground. The 0.22-μF capacitor at the input of the voltage regulator is important and should not be omitted. This capacitor should be placed between the input and common terminal of the regulator *immediately* adjacent to the IC to avoid any chance of instability.

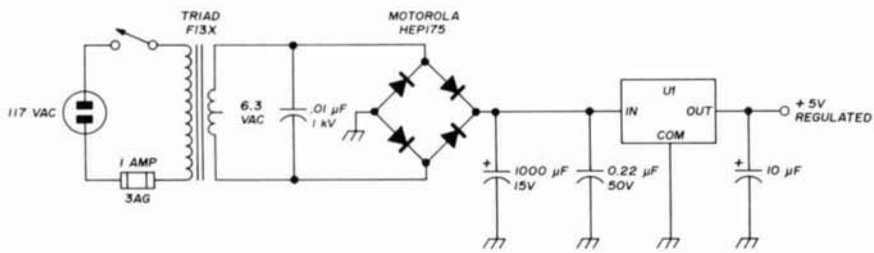
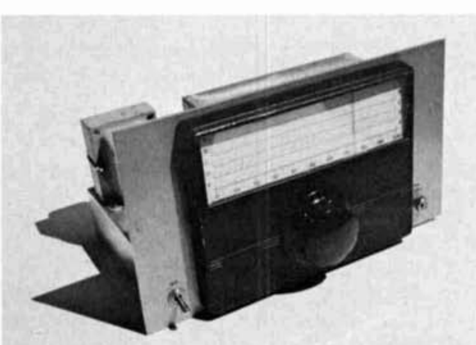


fig. 4. Regulated power supply for the wide-range signal generator. Three-terminal regulator U1 is a Fairchild μA7805, National Semiconductor LM309K or Motorola MLM309K.

The signal generator is built into a 7.5x4.7x3-inch Bud CU-347 cast aluminum box, which is mounted on an 11x7x2-inch aluminum chassis. A 7x12-inch panel is used to mount the Millen 10035 dial assembly. The broadcast variable capacitor is electrically "floating" inside the cast aluminum box and is mounted by screwing it to a Lucite plate, which in turn is mounted on 1/4-inch standoff spacers to the inside bottom of the box. The MC4024P and MC3029P ICs are socket-mounted upside-down on a 2-1/2x2-1/2-inch piece of double-sided copper-clad circuit board. The ICs themselves are not visible, but the socket terminals are conveniently exposed for wiring.

The power supply circuitry is built in the underside of the 11x7x2-inch aluminum chassis. In this way all the parts of the power supply which have large 60-Hz signals on them are well isolated from the MC4024P — which has quite a high modulation sensitivity.

Since the broadcast variable capacitor I used has two 365-pF sections, only one of which is used, it would be possible to add frequency coverage down to 300-kHz by simply adding an spst switch. This was not done in the preliminary model because it was not mechanically convenient. However, such an addition should be considered when building a new version, since the 455- to 500-kHz region is quite useful for i-f alignment.



The wideband signal generator is built into a Bud CU-47 enclosure. Dial mechanism is a Millen 10035.

ham radio

two-stage cavity filter for two meters

Complete construction details for a highly selective resonant filter for 144 MHz

Stirling M. Olberg, W1SNN, 19 Loretta Road, Waltham, Massachusetts 02154

Because of the popularity of the fm mode of communications, the amateur vhf bands are becoming much more active. With this activity comes the attendant equipment problems, which include interference to and from our landmobile service neighbors who, in some cases, use the same geographical location as the amateur station. Overloading the neighboring receiver, or being overloaded, are the most prominent problems. Spurious radiation is another nuisance.

Overloading manifests its presence by the sudden decrease in sensitivity of a receiver which has a signal forced into its input. The overloading signal does not have to be near the operating frequency of the overloaded receiver, but it will be strong enough to get into the frontend and cause the agc to cut down the overall gain of the receiver. Often, when this effect occurs, the operators will not be aware of it because the overload signal bears no intelligence. The reverse of this effect causes problems with the neighboring station in exactly the same manner. If any spurious radiation occurs at the same time you can be sure the amateur is the one to be roasted.

looking for the cure

To improve neighborhood relations with a technician who maintains equipment in the same building and uses the same antenna platform as I do, an investigation was completed which revealed the desensitization of several receivers. One receiver operated in the amateur two-meter fm band and another in the Land Mobile Service on an adjacent frequency allocation. The transmitters for each of these two services were at the 50-watt level.

A probe with a crystal detector was mounted on the tower, halfway between each antenna; meter indicators were located near each of the transmitter/receiver units so that observations of *on time* could be accurately known and used when comparison adjustments were being initiated. It was interesting to note that other services, 10 MHz away in frequency and located geographically on the other side of the hill, were detected and in several instances desensitized the commercial receiver. A plot of the input circuits for the rf amplifier and mixer for the amateur receiver was made. The input circuits and interstage coupling circuits are double tuned and critically coupled by the manufacturer, indicating that previous thought had been given to the matter of high-Q preselection.

A similar test was made on the commercial station receiver. The plot for the amateur receiver preselection circuits, fig. 1, is presented on a scale which clearly shows how the adjacent frequency transmitter could easily control its sensitivity through overloading. To eliminate this problem it was obvious that further selectivity was required for the receiver front-end.

To accomplish this, a major circuit revision would be required. Further investigation of several of the commercial sets revealed that the same problem had been relieved satisfactorily by adding a coaxial filter to the antenna feedlines. These units were simple coaxial tanks, designed with a low coupling coefficient to maintain a high Q, and, therefore, improve selectivity.

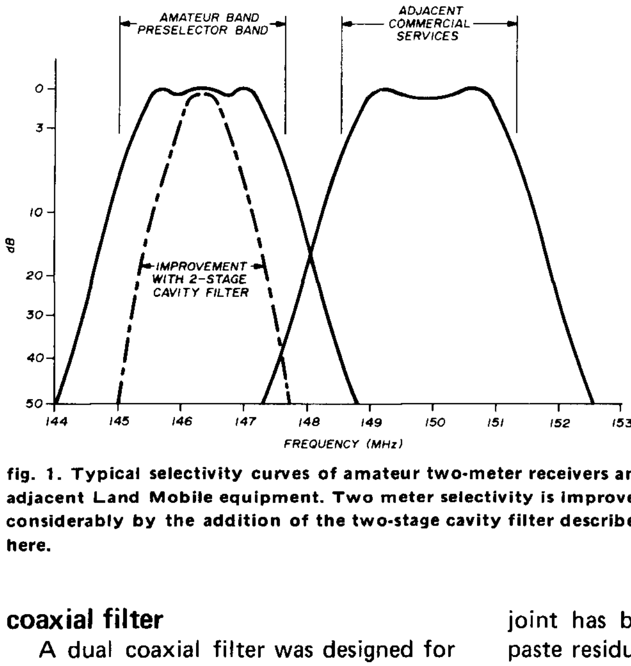


fig. 1. Typical selectivity curves of amateur two-meter receivers and adjacent Land Mobile equipment. Two meter selectivity is improved considerably by the addition of the two-stage cavity filter described here.

coaxial filter

A dual coaxial filter was designed for 145 MHz. The filter was built from plumbing house supplies because these parts are very readily available. Construction details are shown in **fig. 2.** A list of materials is included for 145 MHz which will assist the constructor in locating the required copper fittings (**table 1**).

Assembly of the multiple-cavity filter is simple. First, inspect two 1-1/2x3/4-inch reducing couplings to see that there are no dents on either perimeter. Next, carefully file smooth the lip found in the interior of the 3/4-inch entry. When filing try not to touch the smooth area of the 3/4-inch pipe wall on this fitting, just break down the step so that a piece of 3/4-inch pipe will slide through each fitting.

Lay out the holes to be drilled in the B section of 1-1/2-inch copper water pipe. The lengths of the pipe and hole locations can be determined from the chart accompanying **fig. 3.** Both holes should be concentric; one should be large enough to accomodate the round shoulder on the mounting flange of a SO-239 coaxial receptable. At a point 180° away from the connector hole, a second entry is required which is large enough to allow a piece of 3/8-inch copper water pipe to slip in to a tight fit.

At the base of each reducing coupler drill two holes with a number-28 drill. Slide the B section of 1-1/2-inch copper pipe into the reducing coupler. Align the reducing coupler so that the number-28 holes are parallel with the two large holes in B section. Sweat solder the two parts; use just enough heat to cause the copper to slightly change color. Use soldering paste. When the joint has been soldered, wash away the paste residue with hot water or a cleaning solvent. Try to make the joints as nearly watertight as possible.

The next step requires the addition of part H, a 1-1/16-inch disc which is soldered to the end of part J, a section of 3/4-inch copper water pipe, 17-inches long. Two sections should be prepared. The disc can be a large steel washer or can be cut from sheet copper. It is half of a capacitor used to foreshorten the cavity. It is also part of the tuning system.

Prepare a SO-239 coaxial receptacle by soldering a 3-inch length of number-14 wire to the center conductor terminal. Bend the wire at a right angle directly where it exits the solder point on the connector, insert the end of the wire into the large hole provided for the connector on the B section of copper pipe, and feed the wire down so that it enters the drilled number-28 hole into the reducing coupling. Solder the SO-239 fitting into place. Likewise, solder the wire to the exterior base of the reducing coupling and trim

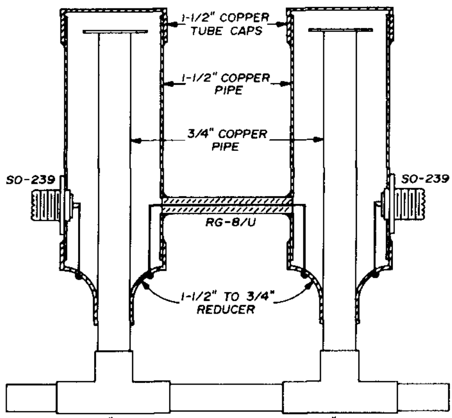


fig. 2. Two-stage cavity filter for two meters is built from common copper plumbing components.

off any excess wire and solder. Align the wire so it is parallel with the pipe wall. This completes the input coupling jack assembly.

output coupling

The next step provides the output coupling link. Prepare an 8-inch section of RG-8/U as follows: strip off the vinyl jacket and the shield braid. Measure 2-1/2-inches in from each end of the center dielectric and cut away the covering to expose the center conductor wire. Bend the wire 3/8 of an inch from the remaining covering, slide the end of the wire into the B section of the cavity through the 3/8-inch hole, and fish it down to the remaining number-28 hole in the reducing coupling. Prepare the second cavity in the same way, leaving out the last step.

Slide a 3/8-inch copper pipe, 3-1/8-inches long, over the coaxial-cable center dielectric and into the hole on the B section of the first cavity wall; carefully solder it in place. Bend the remaining end of the center conductor in the same manner as the opposite end, insert it into the 3/8-inch hole in the second cavity and into the number-28 hole located in the second reducing coupling. The end of the 3/8-inch pipe will now be fitted to the wall of the second cavity and soldered in place. The input/output and inter-cavity coupling are now complete.

The cavity assembly at this point is quite fragile and must be handled as such. Two 1-1/2-inch pipe stands must now be added to each cavity. One stand should be located at the top of each cavity, the other just above the 3/8-inch pipe containing the coaxial coupling element. Fasten the stands to a section of aluminum panel which will serve as a mounting for the filter. Level each pipe stand so that no strain is given to the inter-cavity coupling. If pipe stands are not available, a pipe clamp can be used.

Now, take two 17-inch lengths of pipe with the washer or disc soldered in place and insert the open ends into the 3/4-inch opening of the reducing coupling. Slide each section down to point where the disc is 1/2-inch below the top of the B section.

Be sure to clean these two parts so

table 1. List of materials required for the two-section two-meter cavity filter.

qty	description
2	1-1/2- to 3/4-inch copper reducing coupling (Mueller Streamline style WC400R, part number W1067)
2	3/4-inch drop ear tees, copper (Muller sytle 310, part number A1550)
2	1-1/2" copper tube caps (Mueller style WC415, part number 7013)
1	25-inch length of 1-1/2" copper water pipe, cut into two 12-1/2-inch lengths
1	34-inch length 3/4" copper water pipe, cut into two 17" lengths

The above material can be obtained from plumbing suppliers; part numbers are those of the Mueller Brass Company, a large plumbing parts manufacturer. Similar parts are available from Sears.

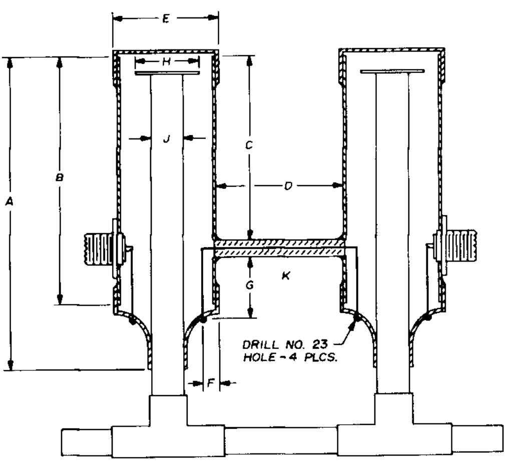
that they are very bright. Use fine steel-wool. Be sure the reducing section is also clean; these pieces will be soldered later and must make a very good connection. Place a tube cap on the open end of each cavity. Both contact points should be clean and bright since these parts also will be soldered. Push the cap down as far as it will go.

tune up

You are now ready to tune up the cavity. It is best to use the transmitter coupled through a vswr bridge to a 50-ohm load for tuneup as shown in the block diagram.

Since the vswr bridge will serve as a resonance indicator, it should be set to the *forward* position. Set the sensitivity control to minimum. When power is first applied to the filters a small indication will be observed. The end of each of the 3/4-inch pipes should be carefully moved in very small steps which will cause the vswr meter to indicate an increase in output. Adjust each pipe until there is no further increase in output level. The sensitivity of the bridge can be adjusted as required. The reflected vswr should be no worse than before the filter was inserted if you follow all the dimensions shown in **fig. 2**.

fig. 3. Construction of the two-stage two-meter cavity filter. Dimensions for other vhf bands are shown in the attached chart.



	frequency MHz			
	50	144	220	440
A	41.0"	17.0"	7.0"	5.0"
B	38.0"	12.6"	5.0"	3.0"
C	31.6"	12.9"	5.0"	3.8"
D	3.0"	3.0"	3.0"	3.0"
E	4.5"	1.5"	3.0"	3.0"
F	1.4"	0.375"	1.0"	0.75"
G	6.0"	2.1"	1.4"	0.75"
H	3.0"	1.06"	2.75"	2.75"
J	0.75"	0.75"	0.75"	1.0"
K	----------see text----------			

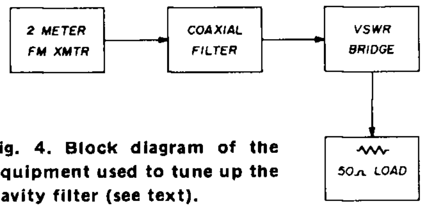


fig. 4. Block diagram of the equipment used to tune up the cavity filter (see text).

If you reverse the input and output connections, there should be no difference from the previous measurements. For this reason, it makes no difference which SO239 jack is used for the input or the output.

To determine the amount of signal loss through the filter simply connect the coax directly to the vswr bridge and note the level. Compare output power without the filter to the level of the output with the filter installed. If you want to determine the ratio of *on* frequency resonance to *off* frequency loss, simply switch the transmitter to the 144-MHz end of the band for the low-end ratio and to 149-MHz for the high end loss. Loss at the high and low ends of the band should be near 40 dB.

When the tuneup adjustments are complete, carefully solder the 3/4-inch pipe to the reducer entry. Solder the top cap in place. The filter is now complete and it can now be installed in the feedline of your transceiver. A set of dimensions for filters for other vhf bands is shown in **fig. 3**.

The improvement at my station has been worth all of the effort and at not too great a cost. I no longer have the desensitizing effect and my commercial neighbor now has a similar filter tuned up on his Land Mobile channel.

ham radio

three-terminal voltage-regulator ICs

Design and construction of regulated power supplies is simplified by the use of three-terminal voltage-regulator ICs

James E. Trulove, WB5EMI, 1409 SW 70th, Oklahoma City, Oklahoma 73159

The new Fairchild 7800-series three-terminal voltage-regulator ICs present some vastly new features not previously available to the amateur. For one thing, they provide a lot of regulation for very little money. However, the use of these regulators takes a new orientation, especially for those readers who have designed and/or built conventional regulated power supplies. In this article, I will cover some of the new aspects of using these IC regulators and will show several recommended circuits for different types of power supplies.

Many new integrated circuits no longer require the addition of several "generalized" circuit elements, such as transistors and diodes, to perform a specific function. Rather, these functional blocks are already combined, within the IC, to perform one very specific function.

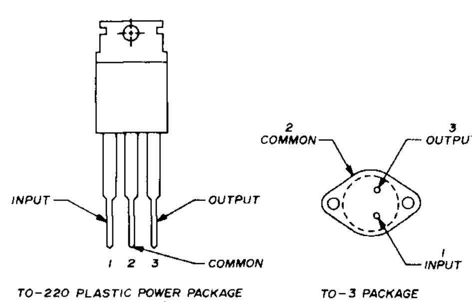


fig. 1. Electrical connections to the Fairchild 7800-series three-terminal voltage-regulator ICs.

The 7800 regulators are good examples of this new breed of IC. Each device in the series is preset to regulate a fixed output voltage. For example, the 7805 is a positive five-volt regulator. A complete list of the 7800 series and the respective preset voltages are shown in **table 1.**

The main advantage of a fixed-voltage regulator is the ease with which it can be used. Since the basic operation of the 7800 IC requires no external components, all you need is a power transformer, a bridge rectifier and a filter capacitor, and you have an instant power supply. You don't have to worry about choosing transistors, biasing them, and protecting the regulator against short circuits.

features

The 7800 voltage-regulator series features a preset voltage tolerance of ±5%, more than adequate for the vast majority of electronics experimenting. The tolerance means that the actual output voltage of an individual 7805 sample, for example, may be anywhere between 4.75 and 5.25 volts. However, the actual voltage regulation, once you have chosen a particular device, is 0.01% per volt, or 0.05% for the five-volt 7805. I doubt that most experimenters need better regulation than that!

Another valuable feature of the 7800 IC regulators is their built-in protective circuitry. The circuit guards against the three most common causes of power supply failures: excess output current, output short circuit and excess heat. The first two causes are listed separately because of the subtlety of a current overload – you may have your project hooked up properly, but are simply demanding a little too much current. The

table 1. Low-cost three-terminal fixed-voltage IC regulators manufactured by Fairchild, Motorola, National Semiconductor and Silicon General.

Fairchild number	National number*	Motorola number	regulated voltage
μA7805	LM340T-5	MC7805	5 volts
μA7806	LM340T-6	MC7806	6 volts
μA7808	LM340T-8	MC7808	8 volts
μA7812	LM340T-12	MC7812	12 volts
μA7815	LM340T-15	MC7815	15 volts
μA7818	LM340T-18	MC7818	18 volts
μA7824	LM340T-24	MC7824	24 volts

*The letter T designates the TO-220 package; for the metal TO-3 package, substitute the letter K. Motorola devices are in a metal TO-220 package.

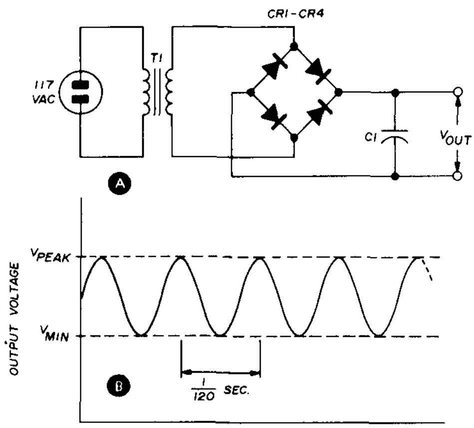


fig. 2. Basic unregulated power supply and typical 120-Hz ripple component.

7800 regulators compensate for these failure modes by internally limiting the output current that can be drawn from the device. In the case of a complete short circuit, only 750 mA, typically, can be drawn from the 7805.

The thermal shutdown protects the regulator from overheating. Additional safe-area compensation of the output transistor prevents the circuit from trying to dissipate too much power. Power capability is 15 watts. This means that you can draw 1 amp of current at 5 volts if the average unregulated input voltage is 20 volts or less, and if adequate heat sinking is provided.

The 7800 series ICs come in two case styles: a TO-220 plastic power transistor case, and a metal TO-3 case. The TO-3, having a lower case-to-ambient thermal resistance, is easier to heat sink, but it is more difficult to mount. Electrical connections are shown in **fig. 1.**

unregulated supply

The unregulated power supply is a

basic element for the properly operating regulated supply. An example of an unregulated power supply is shown in **fig. 2A.** The line voltage is stepped down by transformer T1, rectified by the diode bridge and filtered by the output capacitor. With no load, the output voltage is equal to the peak transformer output

design the unregulated supply. For all the regulated circuits discussed here we will assume a properly designed unregulated supply which can provide 1 ampere of current without allowing V_{min} to drop below the sum of the desired output voltage *and* the regulator maximum voltage drop.

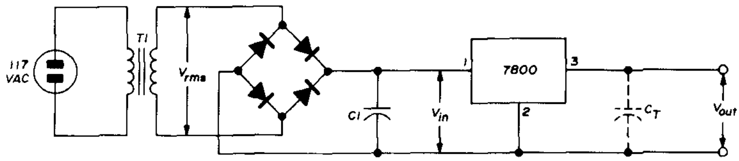


fig. 3. Basic positive voltage-regulated power supply using a 7800 series IC. Capacitor C_T may be added to improve transient response.

voltage (1.4 times the rms voltage), less twice the diode forward voltage drop. As current is drawn, however, the voltage decreases momentarily between the charging peaks of the bridge. This creates the ripple shown in **fig. 2B.**

The average output voltage is between V_{peak} and V_{min}. The greater the load, the more capacitor C1 discharges between charging pulses, and the lower V_{min} becomes, thus accentuating the ripple. For the same load current, the ripple decreases with an increase in the size of C1, up to the point where the bridge can no longer recharge C1 fast enough. One way to eliminate the output ripple is by regulating the output voltage. However, you must never drain so much current as to allow V_{min} to dip below the desired regulated output voltage.

Thus, the first step in building a regulated power supply is to properly

designing the unregulated supply

Although the discussion which follows pertains to the 5-volt 7805, a similar design approach is used with other members of this voltage regulator family. The 7800 series of ICs requires a minimum of 2 volts input-output differential for proper regulation. This means that V_{in}, in **fig. 3**, must be at least 2 volts higher than V_{out}. Since the preset voltage of the output is ±5%, the worst case is 1.05 times the rated voltage, or

$$V_{out\ (max)} = -1.05\ V_{out}$$

table 2. Operating parameters for an unregulated supply designed for use with a 7805 5-volt IC regulator.

Desired operating voltage	V_{out}	5 volts dc
5% tolerance (1.05 V_{out})	$V_{out\ (max)}$	5.25 volts dc
Differential (+ 2 volts)	$V_{in\ (min)}$	7.25 volts dc
Ripple allowance (1.1 $V_{in\ (min)}$)	V_{peak}	7.98 volts dc
Diode drop (+ 1 volt)	V_{peak}	8.98 volts dc
10% line variation (1.1 V_{peak})	V_{peak}	9.88 volts dc
Transformer output (V_{peak}/1.4)	V_{rms}	7.06 volts rms

Since there must be at least a 2-volt differential across the 7800

$$V_{in\ (min)} = V_{out\ (max)} + 2$$

With a 10% ripple, at full current, the peak value of V_{in} should be 1.1 $V_{in\ (min)}$. Since the transformer current must pass through two of the bridge diodes during any charge pulse, two diode drops (approximately 0.5 volts) must be added, $1.1\ V_{in\ (min)} + 1.0$ volt. This is the peak output voltage required from the transformer. If you allow for a 10% variation in the line voltage (105.3 to 128.7 volts for a nominal 117-volt line), you require an extra 10% for the transformer output voltage. These considerations are summarized for a 5-volt supply in **table 2.** In the worst case, the transformer must be able to produce at least 7 volts rms to operate properly.

The only thing that remains is to choose capacitor C1. The capacitor must have sufficient capacitance to prevent its voltage dropping below 7.25 volts when 1 amp is being drawn. A few quick calculations will show that a capacitance of 12,000 μF will meet this condition. Of course, if the actual transformer voltage chosen is greater than the 7.06-volt minimum, a smaller capacitor will do fine, since more ripple can be tolerated.

regulator circuits

The basic hook-up for the 7800 voltage-regulator ICs is refreshingly straightforward, as shown in **fig. 3.** The only embellishment is the optional transient suppression capacitor, C_T, at the regulator output. This capacitor, typically 10- to 50-μF, will improve transient and high-frequency response, but at the cost of increasing output impedance at frequencies below 1 kHz. Additionally, if a battery is used for the unregulated supply, an input bypass capacitor of at least 0.22 μF should be attached across the input to the 7800 (pin 1 to ground).

To increase the current capacity of the 7800s ICs, you may wish to add a pnp series pass transistor as shown in **fig. 4A.** In this application, the pass transistor handles most of the supply current. The 2N4398 transistor shown has a maximum collector current of 30 amperes. The 7800 regulator IC holds the output voltage constant by varying the bias on the

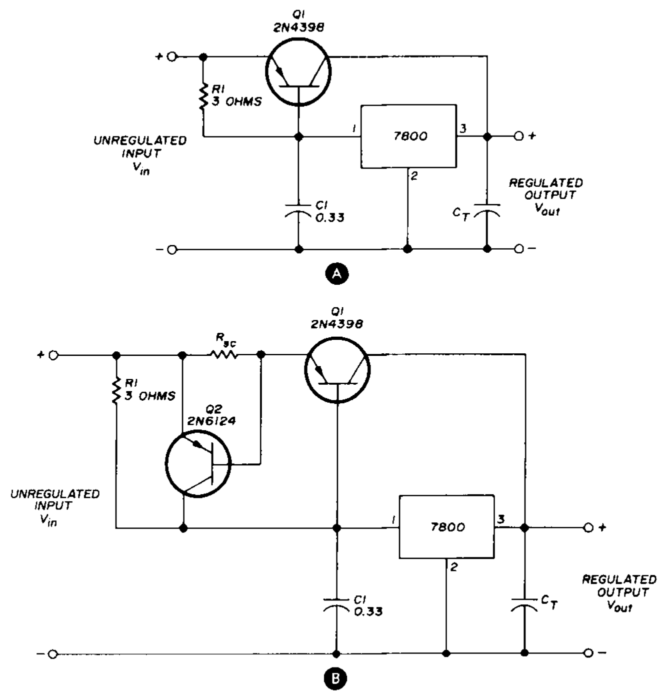


fig. 4. High current voltage regulator circuit using external power transistors. Circuit in (B) includes short-circuit protection for the power transistor (see text).

base of the pass transistor. As the load current increases, the output voltage drops slightly, causing the 7800 to draw more current. This increases the base current of transistor Q1, which brings the output voltage back up by supplying

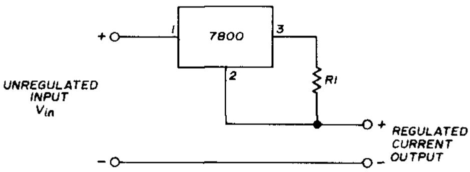


fig. 5. Circuit for using the 7800-series voltage-regulator ICs as current-mode regulators.

more current to the load. Thus, the output current is supplied by both the 7800 IC and the pass transistor.

If you are concerned about a short circuit burning out the power transistor, you can insert a protection circuit consisting of transistor Q2 and resistor R_{sc}, as shown in **fig. 4B**. The 7800 IC protects itself, but has no feedback feature to protect external elements. The series resistor, R_{sc}, may be set for the particular value of current you wish to limit. It should be a small value resistance capable of handling the power through it. Transistor Q2 may be any moderate gain npn transistor which can handle the short-circuit current of the 7800.

The use of this series pass transistor allows you to quickly build a fixed regulated voltage supply for almost any application you may have. In fact, if the protection circuit of **fig. 4B** is used, R_{SC} may be made variable, so as to provide exactly the extent of current limiting needed to protect the load circuit.

Some applications, such as battery charging, require a constant load current, rather than a constant voltage. **Fig. 5** shows the connections to a 7800 for current regulation. The 7800 tries to maintain a constant voltage across R1. In the case of the 7805, the voltage is 5 volts. Obviously, the 7800 will regulate the amount of current through R1 necessary to maintain this voltage. As this supplied current also passes through the load resistance, its current is likewise regulated. In the case of the 7805, five volts across a load of 100 ohms would produce a load current of 50 mA – the charging current for a 500 mA-hr nicad battery. Within the limits of the unregulated input voltage, the 50 mA will be supplied equally well to a simple battery, or a whole stack of them, and the charging current will not change as the cell voltage goes up during the charge cycle.

summary

In conclusion, you should find the 7800 IC voltage regulators to be an invaluable addition to your electronics repertoire. It will free you from the frequent and routine task of building regulated power supplies. Also, you will find yourself using regulated supplies more often because of the simplicity and low cost ($2.20 in single quantities) of the 7800 voltage-regulator ICs.

ham radio

"Remember your application for renewal of your ham license that I was supposed to have mailed four months ago?"

More Details? CHECK—OFF Page 126

low-voltage audio agc amplifier

Description of a wide range audio agc system that operates from a 1.5-volt flashlight battery

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The circuit described here operates from a single penlight flashlight battery with a current drain of 0.5 mA, nominal. It should be ideal as a self-contained unit which can be connected in a microphone cable. The agc control element is a transistor used in the inverted connection to obtain better performance. Those readers who are unfamiliar with audio agc theory and applications are referred to a previous article.[1]

circuit

Fig. 1 shows a schematic of the agc amplifier. Transistors Q2, Q3 and Q4 form a 70-dB voltage amplifier; Q5 is the detector, and Q6 is an emitter follower required to drive the control transistor, Q1.

Resistor R1 and transistor Q1 form a voltage divider which attenuates the signal, as needed, to hold the amplifier output constant. When the input signal is 50 μV or less, the detector has zero output, and Q1 is turned off. As the signal increases, the detector feeds dc to the base of Q1 causing its collector-emitter resistance to decrease; this decreases the signal input to Q2.

With a power supply voltage of only 1.5 volt, the detector must be able to operate from a relatively small peak-to-peak ac voltage, or clipping will occur. The detector shown requires only 0.62 volt peak-to-peak input.

Notice that the control transistor, Q1,

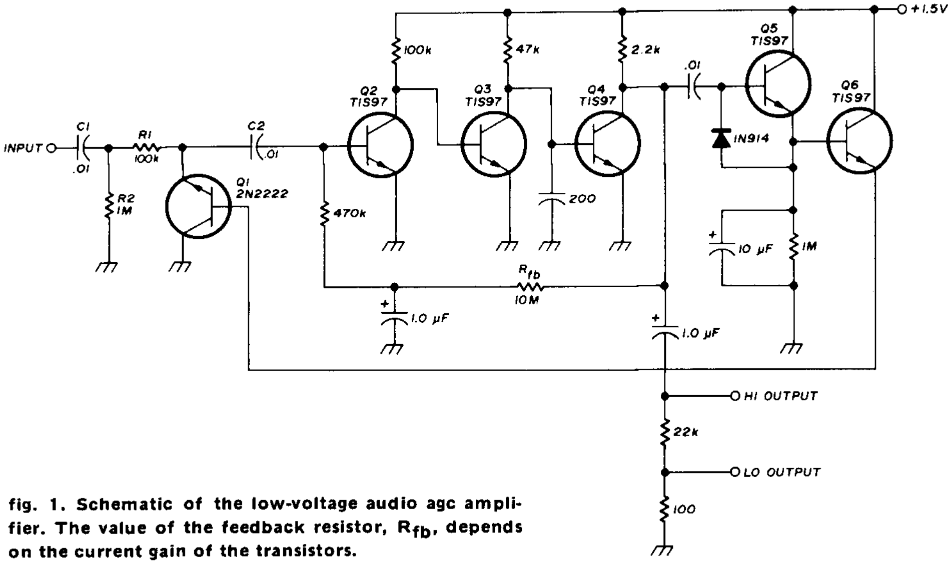

fig. 1. Schematic of the low-voltage audio agc amplifier. The value of the feedback resistor, R_{fb}, depends on the current gain of the transistors.

is shown in the inverted connection as is used with chopper transistors; the transistor is turned on by current flow through the base-collector junction. Current gain is very low in this configuration, being on the order of 1.0, but collector-emitter resistance vs base current is about the same as with the normal connection.

The inverted connection is preferred because it produces a very low offset voltage; offset voltage is the dc voltage appearing between the collector and emitter when the base is biased on with no collector power supply. **Fig. 2** shows how offset voltage is measured. Offset voltage in the normal connection can be 50 mV or more, but it is about 1 mV or less in the inverted connection.

Why is low offset voltage important? When the input signal to the agc amplifier suddenly increases from a very low value to a large value, the detector turns the control transistor on rapidly. If the control element produces a significant dc voltage across its terminals, a transient voltage spike is coupled to the amplifier input. This spike bears no relation to the signal amplitude and can drive the amplifier into hard saturation.

Suppose a 30-mV dc level suddenly appeared across the control transistor. The 70-dB gain of the amplifier would try to amplify the transient to a level of about 100 volts peak. Naturally, the amplifier cannot do this, so it saturates, upsetting the quiescent bias conditions. Recovery time from this strong transient may be one second or more. Under such conditions it is virtually impossible to achieve fast attack times.

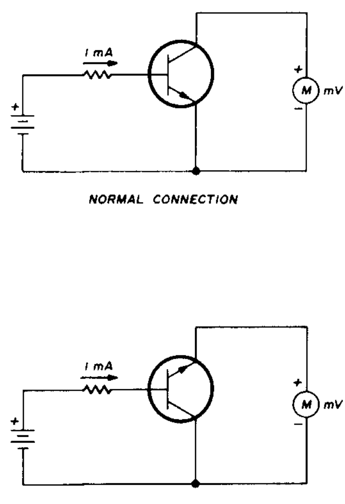

fig. 2. Procedure for measuring the offset voltage in the normal and inverted transistor connection with a millivolt meter connected between the collector and emitter.

Several transistor types were tried in the control transistor socket, but the 2N2222 was the only one that performed well. A transistor designed specifically for chopper applications should work best of all.

50 milliseconds, and release time is about 2 seconds.

conclusion

The results achieved with this circuit show that good performance can be

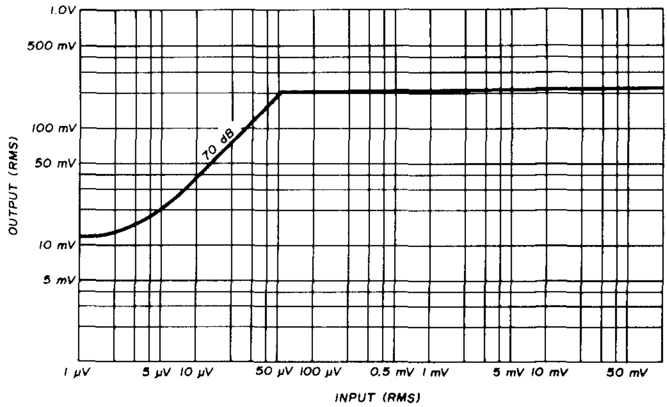

fig. 3. Input/output voltage characteristics of the low-voltage audio agc amplifier.

It is important not to omit the 1-megohm resistor, R2, because it prevents the buildup of dc voltage on the emitter of Q1 due to the capacitive voltage divider formed by C1 and C2.

Two outputs are shown in **fig. 1**; one is the full output of the amplifier, and the other attenuates the output to a level of about 1 mV rms, maximum. The attenuated output should be used when feeding the microphone input of other equipment so it will not be overdriven.

The value of the 10-megohm resistor, R_{fb}, depends on the current gain of the amplifier transistors, and its value may need to be adjusted. The value should be set so that the dc voltage on the collector of Q4 is about 0.8 volt.

operating characteristics

Fig. 3 shows the input vs output voltage curve of the agc amplifier. Although the maximum input voltage shown is 100 mV, inputs up to 1.0 volt rms may be applied without significant distortion of the output. The 3-dB bandwidth of the amplifier is approximately 100 Hz to 8 kHz; attack time is less than

obtained from a 1.5-volt agc amplifier, and that suitable transistors used in the inverted connection for the control element offer improved agc characteristics.

reference

1. C. Hall, WA5SNZ, "Audio AGC Principles and Practice," *ham radio*, June, 1971, page 28.

ham radio

"The last time we went out to eat was because of power failure!"

More Details? CHECK—OFF Page 126

bandpass filter design

How to design image parameter bandpass filters using the lowpass/bandpass analogy

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Very few amateurs attempt to design bandpass or band-stop filters, possibly because of the complex mathematical formulas involved. Most amateurs seem to be unaware of the analogy between lowpass and bandpass filters which can considerably reduce the amount of labor involved in the design of bandpass filters. This analogy has long been known in professional circles[1] but apparently hasn't been described in the amateur literature.

The principal purpose of this article is to describe the lowpass/bandpass (LPBP) analogy. A secondary purpose is to provide some tips on filter design that I have found useful in both amateur and professional applications; specifically, how to design filters using only a reactance slide rule or resonance calculator such as the Shure Brothers or Allied Radio reactance slide rule.

It has been my experience that the biggest difficulty most amateurs find in filter design is simply one of making arithmetical mistakes in computing the parameters — slide-rule errors, plain and simple. When you get the parameters right, the filters usually work! Therefore, anything that will mechanize the calculations or serve to check the calculations is a big help in building filters that work the first time.

The LPBP analogy has the further advantage of giving the filter designer a much better physical insight into the practical constraints imposed on bandpass filters than does a set of cold mathematical equations.

image parameter filters

The type of filters to be discussed are known as "image parameter" filters. Historically, this type of filter was the first to be developed and is entirely suitable for many amateur and professional applications. For completeness I will begin with a brief discussion of image parameter filter design.

A low-pass filter consists of a series of sections as shown in the upper portion of **fig. 1.** As long as all sections are designed for the same impedance level, as many sections as are necessary may be connected in series to give the desired frequency response characteristics.

For convenience, filters are designed on a section basis, the basic section being known as a "prototype" section. After the prototype section has been specified, many different variations are possible, depending on the particular application. The sections are broken out of the composite filter in either of two ways: a *mid-shunt (or pi-section) and a mid-series* (or tee-section). Notice in **fig. 1** that the value of shunt capacitor in the pi-section is one-half that of the composite filter; when two pi-sections are connected together the value of shunt capacitance adds up to the full value of shunt capacitance.

A similar statement can be made for the series inductance of the tee-section. The equations to determine the full value of both the series inductance and shunt capacitance in terms of the cutoff frequency and impedance levels are also given in **fig. 1**. These values are known as the prototype values; one-half of either the inductance or capacitance value must be used in the actual prototype section, depending on whether a pi- or tee-section is used. So as not to complicate the discussion, in the material that follows I will stick with the pi-section.

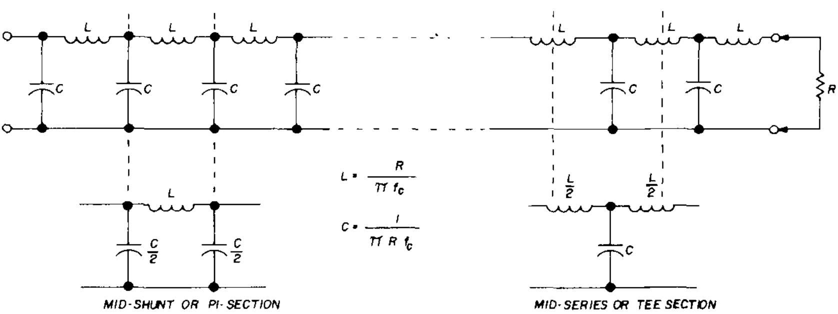

fig. 1. Basic filter sections with prototype design equations. R is the load resistance, f_c is the cutoff frequency.

First, take the equations for the prototype L and C as given in **fig. 1** and put these into the resonant frequency formula to obtain

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{\frac{R}{\pi f_c} \cdot \frac{1}{\pi R f_c}}} = \frac{f_c}{2} \quad (1)$$

You can see that the prototype L and C values should resonate at one-half the *specified cut-off frequency*. This provides the designer with a means of using a reactance slide rule to compute (or check) the prototype parameters.

At this point a numerical example is

useful: Design an audio filter with a 3-kHz cutoff frequency working into a 2000-ohm load resistance. One-half of 3000 is 1500 Hz. Set the reactance rule to 1500 Hz and opposite 2000 ohms read approximately 0.22 henry and 0.054 μF. The prototype section appears as in **fig. 2.**

It is important to realize that the actual resonant frequency of the chosen inductance and capacitance is critical if the filter is to operate as desired. Although some liberties can be taken with the impedance values, the resonant frequencies should be as exact as possible. Therefore, care should be taken to insure that values of inductance and capacitance will resonate at the design frequency, even if the resulting impedance value is not exact. For this reason I suggest that the prototype values of L and C be calculated with a conventional slide rule or calculator, using the formulas, and checked with the reactance slide rule.

At this point a second check point is convenient. I previously pointed out that the resonant frequency of L and C must be one-half the desired cutoff frequency

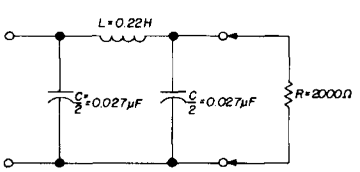

fig. 2. Basic prototype filter section with a 3000-Hz cutoff frequency.

of the filter; because the value of capacitance actually used in the prototype section is one-half that calculated above, the resonant frequency of inductance and capacitance *actually used* in the prototype section will be $1/\sqrt{2}$ or 70.7-percent of the desired cutoff frequency. This is easily verified with the reactance slide rule. In this case 0.027 μF and 0.22 H resonate at 2100 Hz which is 70.7 percent of 3000 Hz; hence, the values check and the complete filter design could proceed with confidence.

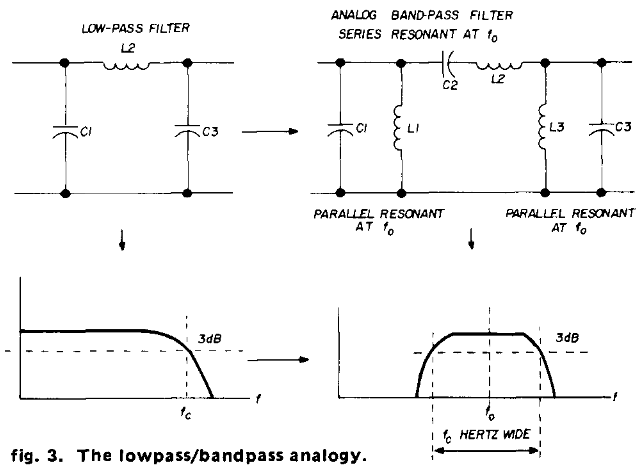

fig. 3. The lowpass/bandpass analogy.

lowpass/bandpass analogy

The LPBP analogy states that to transform a lowpass filter with a cutoff frequency f_c to a bandpass filter with a total bandwidth f_c and a center frequency f_o, it is only necessary to add inductance in parallel with each shunt capacitor so that the combination is resonant at f_o, and to add a capacitor in series with the original series inductance; this combination also resonates at f_o. This is shown in **fig. 3** along with the appropriate response curves.

If there is any confusion at this point, a numerical example should help clear things up. Let's design a bandpass filter centered at 455 kHz with a 15-kHz bandpass such as would be suitable for an fm receiver. The load impedance will be assumed to be 5000 ohms.

First, calculate the inductance and capacitance values for a lowpass filter having a cutoff frequency of 15 kHz:

$$C = \frac{1}{\pi R f_c} = \frac{1}{\pi(5000)(15000)} = 0.00425\,\mu F$$

$$L = \frac{R}{\pi f_c} = \frac{5000}{\pi(15000)} = 0.106 \text{ henry}$$

The basic lowpass filter is shown in **fig. 4.**

A quick check of these values with a resonance slide rule shows that the prototype values, C = 0.00425 μF and L = 0.106 H, resonate at 7500 Hz which is one-half 15 kHz, while the values of inductance and capacitance actually used, C = 0.00212 μF and L = 0.106 H, resonate at 10.6 kHz which is 70.7 percent of 15 kHz. Therefore, we are on a firm foundation and can proceed with confidence.

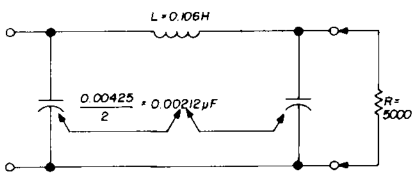

fig. 4. The first step in the lowpass/bandpass analogy — development of the prototype lowpass filter.

The first step in transforming a lowpass section into a bandpass section is to connect an inductance in parallel with the shunt capacitor to resonate at 455 kHz, the bandpass center frequency. This is easily done by using a resonance calculator to obtain a shunt inductance of 58 μH. The second step is to connect a capacitor in series with the series inductance. For this example the capacitance should be 0.001155 μF. The completed bandpass filter is shown in **fig. 5.**

The difference between using the LPBP analog and calculating the individual component values may be compared by considering the bandpass circuit equations shown in **fig. 6.** The interested reader may verify the component values obtained using the LPBP analog by actually solving the equations given in **fig. 6.** The ease of using the LPBP analog will be obvious.

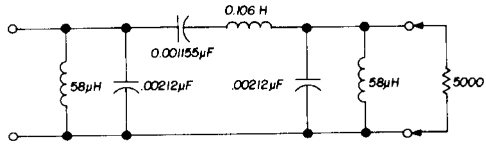

fig. 5. Analog bandpass filter section.

practical filters

The lowpass/bandpass analog also has the advantage that it can give the filter designer a much better feel for the physical realities involved. As an example of this I will attempt to design an exceptionally narrow bandpass filter in terms of center frequency, a 1-MHz wide filter with a center frequency of 50.5 MHz at 50 ohms.

Using the concepts described above, first calculate a lowpass filter with a 1-MHz cutoff frequency at 50 ohms. For a pi-section the inductance will have a reactance of 50 ohms at 500 kHz, or 15.9 μH. The capacitance *actually used* must resonate with this inductance at 70.7 percent of 1 MHz or 707 kHz. This is 0.00319 μF. So far, so good; the lowpass filter is shown in **fig. 7.**

The next step is to transform the lowpass filter into a bandpass filter. Now comes the joker which is easily seen from the LPBP analogy. To make this transformation it is necessary to resonate the 3190-pF shunt capacitor to 50.5 MHz. Now, 3190 pF is a lot of capacitance at 50.5 MHz; it is impractical to resonate

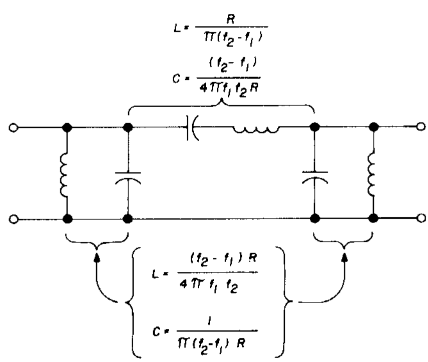

fig. 6. Design equations for the bandpass filter section. Frequency f1 is the low-frequency cutoff, while f2 is the high-frequency cutoff.

this much capacitance with any reasonable inductance. Resonating the series inductance to 50.5 MHz presents no serious problem and requires about 6 pF.

Thus, you can see from the LPBP analogy why a filter of this type is not practical in this application; the bandwidth is too narrow in terms of the center frequency. Before giving up, however, let's try a tactic frequently used by filter designers when one or more of the components turns out to be an impractical value — change the impedance level.

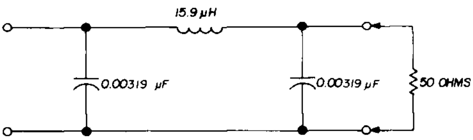

fig. 7. First step in the development of a 1-MHz wide bandpass filter centered at 50.5 MHz.

If the impedance level is increased to 5000 ohms, a factor of one hundred, the shunt capacitors decrease in value to 31.9 pF and the shunt inductances become 0.31 μH, both of which are at least in the ballpark of being practical. The value of series inductance increases, however, to 1590 μH, requiring only about 0.06 pF to resonate at 50.5 MHz. This is an impractically small value.

Therefore, it appears that a filter centered at 50.5 MHz with only a 1-MHz bandwidth is impractical at any impedance level, and other types of filters, such as coupled tuned circuits, must be used to obtain the desired selectivity. These filters, however, are beyond the scope of this article.

The preceding example shows how the LPBP analogy gives the filter designer a much better feel for the practical problems involved than does a purely mechanical application of the design formulas. The 455-kHz bandpass filter with a 15-kHz bandwidth had a lower ratio of center frequency to bandwidth, and also operated at a considerably lower frequency so that component parameters were much more realistic.

summary

After reading the above material you may ask, "Does the LPBP analogy work in the opposite direction; i.e., is there a bandpass to lowpass analogy?" The answer is, "Yes, provided all tuned circuits are tuned to the same center frequency."

It should also be noted that the LPBP analogy described here gives only one particular class of bandpass filter. There are many other types of bandpass filters, the most notable of which is probably a series of tuned circuits coupled by means of capacitance or inductance. The design of coupled tuned circuits is a subject in itself.

Although I have used an image parameter designed lowpass filter as the starting point in this article, the analogy applies equally well to lowpass filters designed on a Butterworth or Tchebyscheff basis.

reference

1. Vernon D. Landon, "The Band-Pass/Low-Pass Analogy," Proceedings of the IRE, December, 1936, page 1582.

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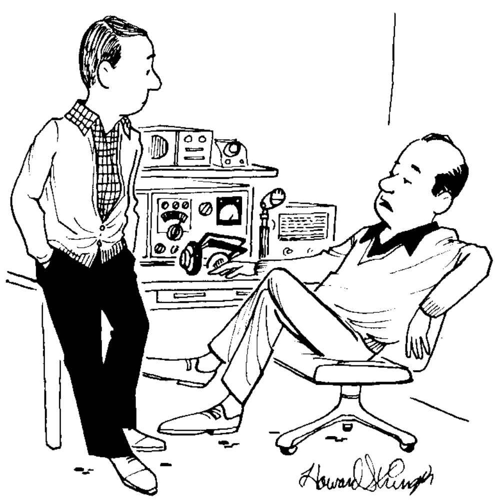

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More Details? CHECK—OFF Page 126

introduction to the digital mixer

How to use the D-type flip-flop IC as a frequency mixer

Gerd H. Schrick, WB8IFM, 4741 Harlou Drive, Dayton, Ohio 45432

Basically, a mixer can be thought of as a switch operating at one frequency which will or will not pass a signal at another frequency. A signal interrupted in this manner generates a combination of various new frequencies with the difference- or intermediate-frequency as the desirable output. Normally, this desired frequency is filtered out for further processing.

In practice the requirements for a mixer are a nonlinear switching device and a large injection signal. These two requirements complement one another to a certain extent since a very nonlinear device requires less oscillator injection, while more injection is required with less mixer nonlinearity.

When working with digital circuitry you are dealing with two voltage levels; therefore, all signals must be square waves or close to it. This means that if you use a smooth sinusoidal signal, it must be converted to a square wave with a circuit such as a Schmidt trigger. If the sinewave is large enough, a simple diode clipper will do the job.

The digital equivalent of a simple frequency mixer is a gate with two inputs such as the 7400 IC. The output contains all the frequencies because the gate does not have a memory and follows momentary changes of either frequency; the desired output frequency must be filtered out.

If you use an edge-triggered D-type (delay) flip-flop such as the 7474 as a mixer, the leading edge of the square-wave oscillator pulse transfers the input signal to the output, and the output remains at this new level until the next oscillator pulse samples the input signal as shown in **fig. 1.** When the oscillator pulse is out of step with the input signal it turns off the output. Thus, the output is a square wave at the intermediate frequency which needs no filtering, except possibly to remove the odd-order harmonics.

There are several D-type flip-flops which can be used for this application. The common TTL 7474 can be used up to about 25 MHz. The high-frequency version of this IC, the 74H74, is usable up to 43 MHz. The Schottky TTL version, the 74S74, can be operated to 100 MHz. For even higher frequency use, Motorola has introduced the new MECL MC12000 digital mixer, which is a D-type flip-flop which can be used up to 250 MHz. The MC12000 has built-in logic converters so it can be interfaced directly with slower TTL 74-series ICs.

Since this mixing scheme is basically a sampling technique, the ratio of the two frequencies to be mixed cannot exceed 2:1. That is, if the input frequency is lower than the oscillator frequency, it cannot be less than half; if the input frequency is higher, it cannot be higher than twice the oscillator frequency.

A typical digital mixer circuit using TTL ICs is shown in **fig. 2.** In this circuit a 7400 TTL gate is operated as a crystal oscillator. The other two gates of the 7400 are used as input buffers to the mixer, a 7474 D-type flip-flop. In this circuit the rf input signal must be lower than the crystal frequency, and the i-f signal must be less than half the crystal oscillator frequency.

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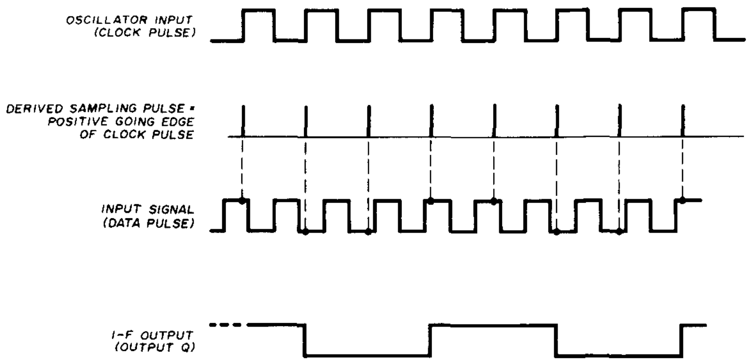

fig. 1. Waveforms in the digital mixer. Information from the input signal (data pulse) is transferred to the output by the positive going edge of the oscillator pulse (clock pulse). When the oscillator input is at either the high or low level, the signal input has no effect.

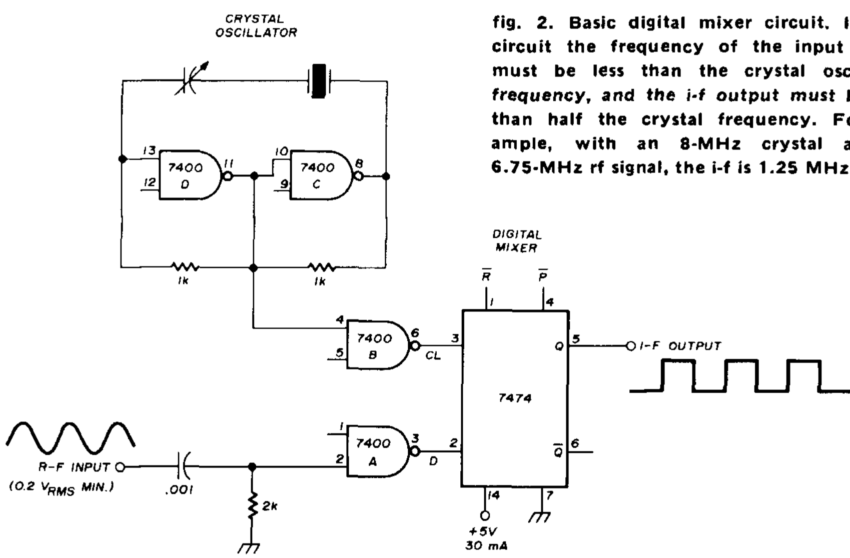

fig. 2. Basic digital mixer circuit. In this circuit the frequency of the input signal must be less than the crystal oscillator frequency, and the i-f output must be less than half the crystal frequency. For example, with an 8-MHz crystal and a 6.75-MHz rf signal, the i-f is 1.25 MHz.

narrowband modifications for the Regency HR-2 series of vhf-fm transceivers

How to install the Regency narrowband kit in the popular HR-2 series of two-meter fm transceivers

Paul Dobosz, WA8TMP, 2500 S. Hickory Ridge Road, Milford, Michigan 48042

The extremely popular HR-2 series of two-meter fm equipment introduced by Regency in 1970 has become an amateur favorite. At the time of introduction the desirability for an extremely selective narrowband receiver was not evident. This prompted Regency to build the HR-2 as a wideband unit.

As an increasing number of repeaters go into operation the wideband fm transceiver is plagued by annoying adjacent-channel interference. In many metropolitan areas all the repeaters are operating narrowband, in and out (deviation of ±5 kHz). With the large number of repeaters and growing popularity of two-meter fm, narrowbanding to conserve operating space is imperative.

The entire Regency family of HR-2 transceivers (HR-2, 2A, 2S and 2MS) can be easily modified for narrowband operation. Narrowbanding the transmitter is accomplished simply by setting the deviation control for a peak deviation of ±5 kHz. Narrowbanding the receiver requires some new parts and a few simple adjustments.

Before installing this modification in my receiver, I was plagued with adjacent-

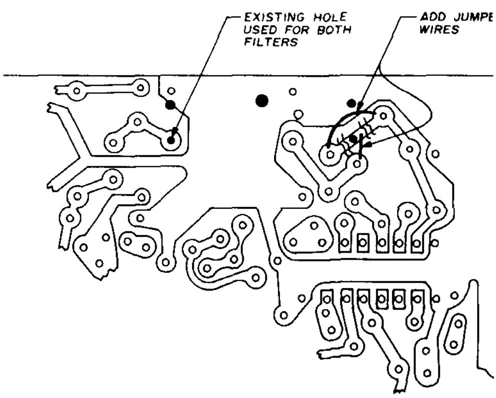

fig. 1. Bottom view of the HR-2 series receiver i-f circuit board, showing modifications required before installing the Regency MA-46 modification kit. One circuit trace must be cut in two places. The black dots indicate new holes for the new, narrow-band ceramic filter supplied with the kit of parts. A drilling template is shown in fig. 2.

channel interference on 146.76 MHz from the local repeater on 146.79 MHz. It was impossible to copy anything on 146.76 when the local repeater was transmitting. This repeater has an effective radiated power of more than 60 watts and is located less than a mile away from my station. With the circuit modification MA-46 Narrow Band Filter (70-dB) Modification, the i-f board is identical to the one Regency uses in their FCC type-accepted marine and fm business-band equipment.

If you are the owner of a HR-2A, 2S or 2MS the instruction sheet provides all the information you need to install the

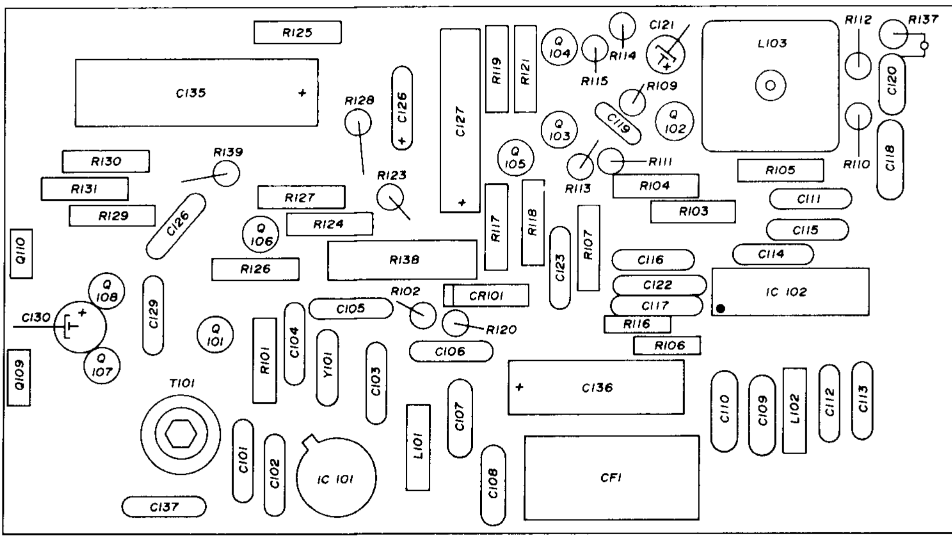

fig. 3. Parts layout of the Regency HR-2 series receiver i-f board. R133, R134, R135, R136, C134 and CR102 are located on the bottom of the board.

it is impossible to tell if the local repeater is operating or not while listening on 146.76 – even when the squelch is turned off.

modification kit

The basic kit of parts necessary to modify the HR-2 series transceivers is available from Regency. The kit consists of a new higher quality ceramic filter with an extremely steep selectivity curve that virtually eliminates adjacent-channel interference, two shielded coils to replace unshielded ones originally supplied with the rig, three capacitors to adapt the i-f circuit to the new ceramic filter, and two resistors to change the sensitivity of the noise-operated squelch to match the new filter. With the addition of the kit, the kit because Regency uses the same i-f board in most of their units; all the holes and spaces for the new additional parts and the holes for the new narrowband filter are already there.

HR-2 modifications

If you are one of the many people who own the original HR-2 fm trans-

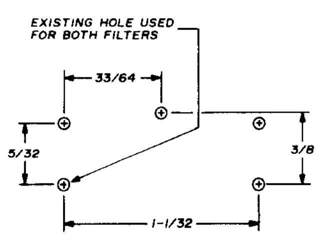

fig. 2. Drilling template for the new narrow-band ceramic filter (foil side of board).

ceiver, don't despair. With a little change in the circuit board it is possible to add the modification kit to the earlier models. When the job is finished the i-f board will be electrically identical to the HR-2A.

Circuitwise, the HR-2 and HR-2A are nearly the same in the modification area, but a different circuit board and different parts numbers complicate the instructions supplied with the Regency MA-46 modification kit. The instructions furnished with the kit should be followed, except as noted here. The parts layout on the reverse side of the instruction sheet should *not* be used. Instead, use the information in **fig. 3.**

changes to MA-46 instruction sheet

1a. Remove the old ceramic filter, CF-1

b. Perform the modifications shown in **fig. 1** to the i-f circuit board (301-528-B) and drill the holes for the new ceramic filter using the dimensions given in **fig. 2,** using the existing hole indicated to locate the new holes. Earlier models may have two resistors (R133 and R134) soldered to the foil side of the circuit board as shown in **fig. 4.** If R133 and R134 (both 6.8k) are present, remove them as they are no longer required.

c. Mount the new ceramic filter.

2. Replace the following capacitors with the values indicated.

a. Replace C108 with a 390-pF capacitor.

b. Replace C109 with a 270-pF capacitor.

3. Add C110, a 250-pF capacitor. In early models this capacitor may already be installed, but to assure the correct value, replace any existing C110 with the capacitor furnished in the MA-46 kit.

4. Replace the following resistors with the values indicated

a. Replace R111 with 5.6k resistor.

b. Replace R112 with 2.2k resistor.

c. Replace R137 with a 100 ohm, ¼-watt resistor (not furnished). R137 may be missing on early models. If it is missing it must be added. R137 is located just forward of L103 and installed vertically, as shown to the right (electrically, R137 is connected between C120 and the emitter of Q-102).

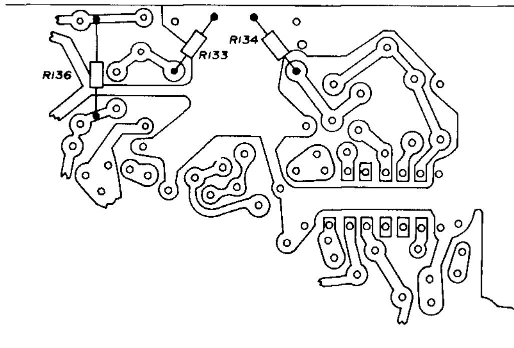

fig. 4. Foil side of the HR-2 receiver i-f circuit board showing the placement of R133, R134 and R136. (R133 and R134 are only present in early models of this fm transceiver.)

d. In early versions of the HR-2, R136 was omitted. In later models it was located on the foil side of the circuit board as shown in **fig. 4.** R136 is a 22k, ¼-watt resistor. If R136 is missing it should be added to the foil side of the circuit board as shown in **fig. 4.** Electrically, R136 is in parallel with L101.

5. Follow the instructions furnished with the MA-46 kit from **step 5** thru to the end of the instruction sheet

With the addition of the MA-46 modifications described here the performance of the Regency HR-2 family of fm transceivers is as good as the latest fm equipment. Furthermore, it can be obtained without the expense of a new rig. The MA-46 modification kit is available from Regency for $22.50, not a bad price when you consider it's almost like getting a brand new receiver, free of that adjacent-channel interference that used to be so annoying.

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More Details? CHECK—OFF Page 126

simple high-gain wire antenna for high-frequencies

Design and layout of a four-element, double-extended Zepp that provides up to 7-dB gain on 15 meters

Alvan L. Mitchell, W6QVI, 24765 Calaroga Avenue, Hayward, California 94545

There's an old saying that you can't get something for nothing, especially when you're working with antennas, but you *can* make one wire antenna, the length of a 75-meter dipole, work like a bomb on 75 and deliver 7-dB broadside gain on 15! This is only one-half dB less than a three-element beam on this band. I call the antenna the FEDEZ – Four-Element Double-Extended Zepp.

Many amateurs have used the extended double Zepp which gives 3-dB gain at its design frequency. However, with the addition of phasing stubs and two more elements you can obtain up to 4-dB more gain. All it takes is a little arithmetic which, in my case, was supplied by W6DMY. The basic design was taken from the 1943 edition of the ARRL *Antenna Handbook.* The dimensions for any frequency are given in electrical degrees in **fig. 1** (remember that 180° = 1/2 wavelength).

Since most of my on-the-air activities are confined to nets on 75 and 40 meters, with hamming just for fun on 15, the four-element double-extended Zepp I use has a 21.3-MHz center frequency (see **fig. 2**).

Although the two 7.68-foot phasing stubs can hang straight down from the antenna as shown in **fig. 2,** I use lumped constants for the two outer stubs as shown in **fig. 3.** Part of the 450-ohm open-wire feedline is used as the center phasing stub. Each of the lumped-constant stubs I use consist of an 11-turn coil, 2-inches in diameter, 2-3/4 inches long, wound with number-12 wire. Each end of the phasing coil is supported by the strain insulator as shown in **fig. 3.**

With this antenna I have yet to receive less than an S9 report on the SARO Bourbon net that meets every night on 75 meters, especially from San Diego and Medford, Oregon. On 15 meters I have received numerous S9 reports from the East coast as well as from Japan. WØQWH in Stanley, Kansas, who has given me signal checks on 47 different antennas over the past year, gave me his

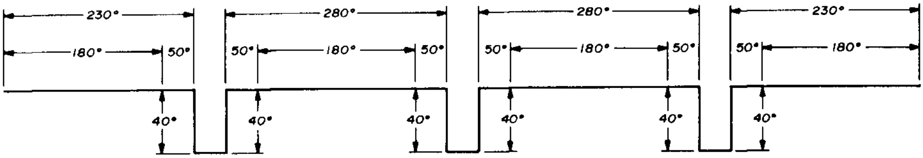


fig. 1. Basic design of the four-element, double-extended Zepp antenna. All dimensions are given in electrical degrees (180° = 1/2 wavelength).

best report, although it wasn't S9 — he apparently has a very stingy S-meter!

The dimensions of my urban lot require that I use this antenna in the inverted-vee configuration. This detracts from the gain somewhat because the wide spacing between the centers of the elements determines gain, and the drooping legs reduce this distance slightly. However, since I feed the antenna with 450-ohm open-wire ladder line to the Ultimate Transmatch,[1] I think I've at last found the ultimate antenna to go with my ultimate transmatch. I don't think you can beat it for city-sized lots.

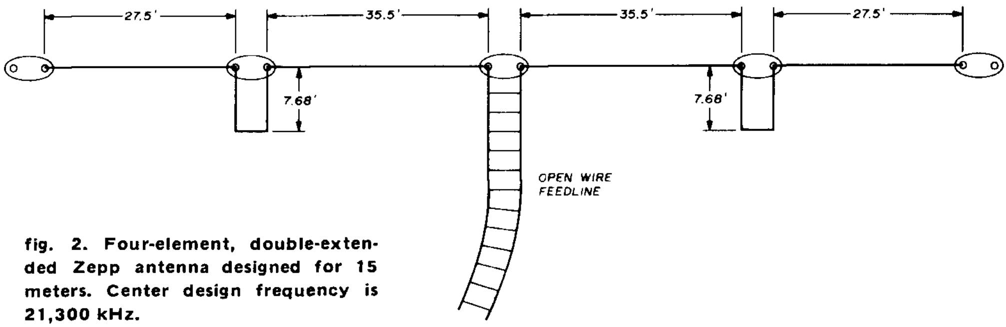


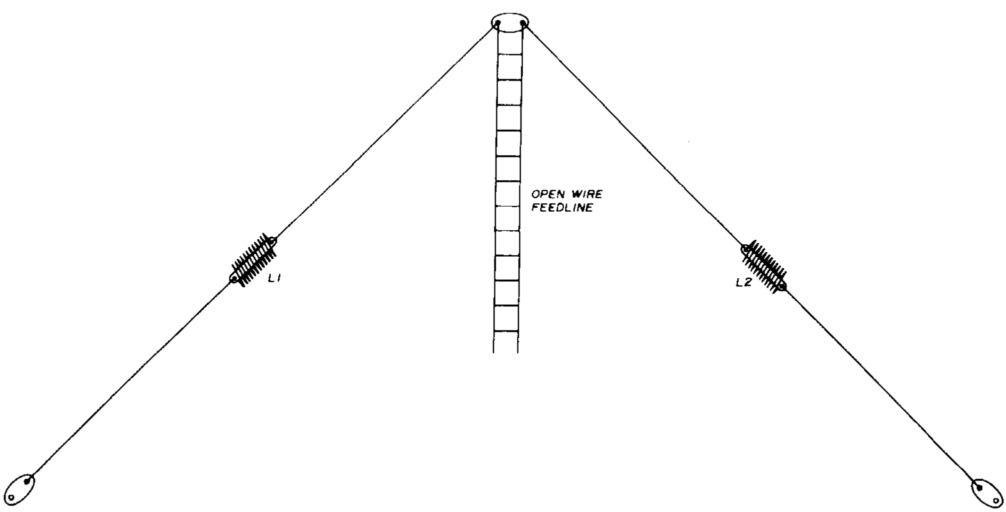
fig. 2. Four-element, double-extended Zepp antenna designed for 15 meters. Center design frequency is 21,300 kHz.

reference

1. Lewis G. McCoy, W1ICP, "The Ultimate Transmatch," *QST*, July, 1970, page 24.

ham radio

fig. 3. The two outer phasing lines can be hung down from the antenna as shown in fig. 2, or phasing inductances may be used as shown here. L1 and L2 are each 11 turns no. 12, 2" diameter, 2-3/4" long. Antenna may be used in the inverted-vee configuration if space is limited.



feedpoint impedance, characteristics of practical antennas

A discussion of antenna feedpoint impedance, and the effects of the resistance and reactance components in practical antennas

Carl C. Drumeller, W5JJ, 5824 N.W. 58 Street, Warr Acres, Oklahoma 73122

The feedpoint impedance, the radiation resistance and dissipative resistance, and the reactance of a common dipole antenna are matters that need clear understanding if you are to inquire deeply into the functioning of that indispensable component of a radio station: the antenna. The purpose of this article is to define and to describe the Z, R and X of antennas, not in a highly technical manner but simply and with only enough detail to distinguish one from another and to show the role each plays.

First, let's consider a center-fed dipole antenna, one a half-wave long (electrically) at the operating frequency, and one out in the clear far enough to have a very minimum modification of its normal characteristics by the influence of its environment. Textbooks tell us that such an antenna will have a feedpoint impedance (Z_f) of 73 ohms, and that this impedance will be purely resistive (no reactance). In the real world, such an antenna seldom is found!

Let's deal first with the ideal dipole, then with the real. In the ideal dipole, Z_f, the feedpoint impedance, will equal R, the composite of the radiation resistance and each of all of the dissipative resistances. These dissipative resistances include the ohmic resistance of the antenna, insulation losses, dielectric losses and absorption losses. These are easy to visualize. You know that the antenna wire has resistance, even though it's made of highly-conductive copper. You know that no insulator is perfect; so even the best has some loss. You know that somewhere within the near-field of the antenna there must be an insulating object that introduces dielectric losses, however small. And you know that somewhere within the near-field there must be some material that will absorb radio waves.

radiation resistance

Radiation resistance, though, is a different matter! In the first place, it's not a true resistance. It acts like a resistance in some ways, but not in every manner. For instance, a real resistance, when radio-frequency current flows through it, converts the electrical energy into heat, another form of energy. Radiation resistance doesn't do this.

What, then, does it do? Nothing! It's just a term which describes an attribute of an antenna, an attribute which bears a superficial resemblance to a real resistor.

The need for such a term comes about from the fact that all of the rf power that flows into an antenna doesn't get converted into heat. Some (and, we hope, a greater part) of the rf power is radiated out into space. It's convenient to speak of an antenna's characteristics as if all of the rf energy fed into it were dissipated just like that portion which produces heat. To make this fiction plausible, we assign an imaginary resistor to the antenna and call it "the radiation resistance."

When we put a known amount of rf power into the antenna, defining it as $W=I^2R$, and having a known amount of current, we have a large enough value of resistance to make the formula valid. We've taken care of not only the amount of power that was dissipated in the various real resistances and equivalent resistances but also the amount of power radiated into space; the latter being equal to what a real resistance of a value the same as the radiation resistance would have dissipated in the form of heat.

Let's run that through again. Taking a purely imaginary situation, let's conjure up an antenna that has only real resistance, a real resistance of one ohm, and feed one ampere of rf current into it. According to the formula, only one watt of power is going into that antenna, and all of it is being converted into heat with none of it being radiated.

Now, conjure up another antenna with one ohm of real resistance and 49 ohms of radiation resistance. With the same one-ampere of rf current going into it, the formula tells us that 50 watts of rf power is going into the antenna – one watt is squandered as heat and 49 watts are radiated. Quite an improvement!

This brings us to a cardinal rule: Make the ratio of radiation resistance to dissipative resistance as high as you can. This is not too difficult to do with a half-wave or even a quarter-wave antenna, but when you attempt it with a really small antenna, say a tenth-wave, you run into a real problem. That's why the engineers who design 80-meter mobile antennas work up such a sweat over their drawing boards.

So much for radiation resistance. Just remember that it's an imaginary resistance that accounts for the power being radiated by the antenna.

antenna reactance

Now for the reactance. Remember, we started out with an ideal antenna, one that was resonant and therefore resistive. It might be resonant on, say, 7,257,376 Hz, but when you breathe on your transmitter and it drifts to 7,257,377 Hz, the antenna departs ever so little from resonance. As it departs from resonance, it loses that purely resistive status. If the frequency goes higher, a bit of inductive reactance is introduced; if it goes lower, the introduced reactance is capacitive.

Just how much the antenna departs from resistive to resistive-plus-reactive status, or, rather, the rate at which it departs for a given change of frequency, depends upon several factors. For the simple dipole we're considering, the chief of these factors is the antenna's diameter to length ratio. The larger the diameter of the radiator for a given length, the less reactance introduced for a given change of frequency. The almost-obsolete cage antenna merits much consideration, for it gives a very favorable ratio of diameter to length.

When reactance is present the feedpoint impedance, Z_f, no longer equals R. It is given by

$$Z_f = V\sqrt{R^2 + X^2}$$

with the R still the grand total of all the resistances (radiation, ohmic, etc.) and X either inductive ($+X_L$) or capacitive ($-X_C$), as the case may be. In either case when X is squared it's a positive value, so forget the sign.

There's one thing you musn't forget, though. That's the matter of reactance not being able to absorb power. Ponder this, for it's quite important! Think of what it involves. The feedpoint impedance may go high and you feel that the dissipative resistance of your antenna is low. You rejoice, believing you're radiating more power, a valid assumption only if it were the radiation resistance that was going up. You can't make a purely-reactive termination accept power. One that's partly-reactive and partly-resistive, yes. One that's purely reactive, no.

Don't jump to the conclusion that reactance in an antenna is an evil thing. In certain antenna designs it plays a vital role, but this is not an article on antenna design. If you want to look into that subject, get a reliable textbook, preferably one written by Kraus, LaPorte, or some other recognized authority on the subject. There's a wide difference between the simple dipole we're discussing and a complex antenna. For this article we'll stick to the dipole!

If your dipole is reactive to a degree, as are the vast majority of such antennas, don't worry about it. If it does give you concern, remember that the reactance can be cancelled out by the introduction of an equal and opposite reactance. For example, if the antenna exhibits 10 ohms capacitive reactance, this can be negated by introducing 10 ohms of inductive reactance. This conjugate reactance can be placed at the feedpoint of the antenna or at any point between that feedpoint and the active device in your transmitter. Its position doesn't matter so long as it reflects that conjugate reactance into the antenna. Keep in mind that the resistive component of the antenna's impedance, which will not be affected by these manipulations to cancel reactance, is going to accept the rf power.

resistance transformation

The resistive component can be transformed by many and various means to any convenient numerical value that you might elect to stipulate. Again, this can be done at the feedpoint of the antenna or at any place between that feedpoint and the active device in your transmitter.

In each instance, there is some slight advantage in having the transformation take place at the antenna's feedpoint. With some transmitters, ones poorly designed or manufactured to meet a price and not to provide quality, it is imperative that the transformation take place between the antenna's feedpoint and the transmitter's antenna terminal. This, though, is strictly a transmitter deficiency.

summary

To sum up, the feedpoint impedance of an antenna is a complex quantity, consistituted by both resistive and reactive components. The resistance component may be made up of many constituents. Of these, one, the radiation resistance, is not a true resistance but an imaginary one invented to account for the rf energy radiated by the antenna. The several other constituents of antenna resistance are all dissipative in nature and should be held to a minimum in design. Radiation resistance should be high as compared to the total of the other resistances. Some element of reactance is present in most antennas, but this is not a significant deficiency and may even be used to advantage in some designs.

ham radio

improved logic test probe

This improved logic test probe checks binary levels as well as pulse coincidence

R.H. Fransen, VE6RF, 227 Cottonwood Avenue, Sherwood Park, Alberta, Canada

Since I have always been interested in test equipment, the TTL logic probe with a built-in memory described in a recent issue of *ham radio*[1] proved very interesting. I made some changes to the basic circuit so that it can take the place, in many instances, of an item we would all like to own but can't afford, a dual-trace oscilloscope. The design uses three ICs, some additional switches and more hardware.

Since I wasn't able to obtain some of the parts used in the original logic probe, like good, bright LEDs, some circuit changes were made as needed. The completed unit may look a little clumsy in its mechanical design because I used what was available, but the probe does the job it's supposed to do, and that's what counts. If you have access to better materials you can dress it up any way you like.

the circuit

The logic probe circuit, **fig. 1,** has two inputs, *main* and *auxiliary.* In the off position of the *off-aux* switch the unit operates as in the original design. However, in this circuit you can switch the memory off with the *off-mem* switch so you don't have to keep pushing the button when using the probe as a binary level indicator.

In the *aux* position of the *off-aux* switch two inputs are needed at the same time. The level of the pulse into the *aux* jack is selected by the *aux* + or - switch (see **fig. 2**). To check the coincidence of pulses, just connect a patch cord from the *aux* jack to the second point on the logic circuit you are checking, and the probe will indicate it.

The parallel RC circuit in series with the *aux* input is to protect the probe against a direct short to common in case the *aux* input is connected and the *off-aux* switch is in the *off* position. The

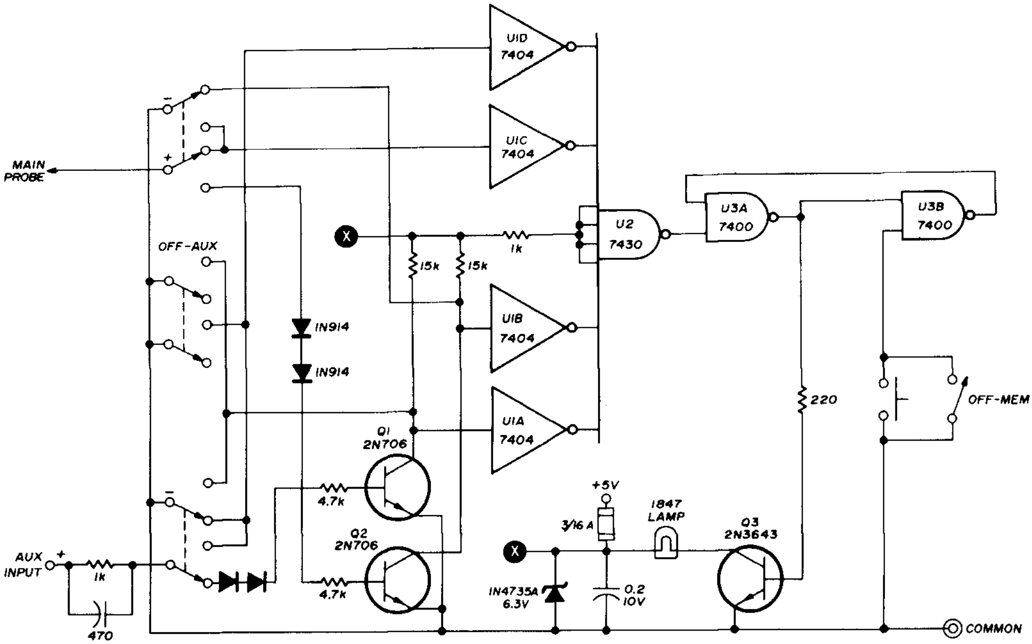


fig. 1. Circuit for the improved logic test probe. All signal diodes are 1N914, all resistors are ½ watt.

470-pF capacitor prevents too much pulse slow down. The 1N914 diodes serve to bring up the high trigger threshold voltage to prevent noise triggering.

I was unable to obtain a decent, bright LED, so I used a long-life number-47 bulb with a switching transistor. At the voltage used, the bulb should last forever, and it's still bright enough to be seen, even in bright sunlight.

The *common* of the TTL circuit is connected to a pin jack for those cases where the circuit under test cannot handle the probe current requirements. With the 1847 bulb the probe needs a total of about 160 mA; changing to a LED would cut probe current to 60 mA.

construction

The entire logic probe circuit is built into a 3¼x2¼x1½-inch aluminum box similar to a Minibox, and fed by a small audio-type coaxial cable with alligator clips on the opposite end. The main probe is mounted on a small ceramic stand-off insulator. The 0.2-μF capacitor is mounted on a small section of Vero board right where the feed cable is connected. The Vero board with all the ICs and other components is mounted on the inside of the aluminum box. All the ICs are mounted in sockets. (It is especially important to mount the 7404 IC in a socket since it serves as a cheap fuse and will burn out if too high voltages are applied to any of the probe inputs.)

operation

In use, the metal case of the logic probe is left floating. Supply current with the lamp off is 26 mA; with the lamp on, current drain is 160 mA. *Main* trigger threshold voltages are +1.5 volts (high) and +1.3 volts (low). *Auxiliary* trigger threshold voltages are 1.5 volts (high) and 0.7 volt (low). The *aux* input can be used by itself if the *main* input is switched to minus (−) and connected to *common.* This may be useful at times since the low level of the *aux* input is half as low as the low level on the *main* input.

If a separate power supply is used for the logic probe only the *common* of the probe must be connected to the negative line of the TTL circuit under test. It should also be kept in mind that when

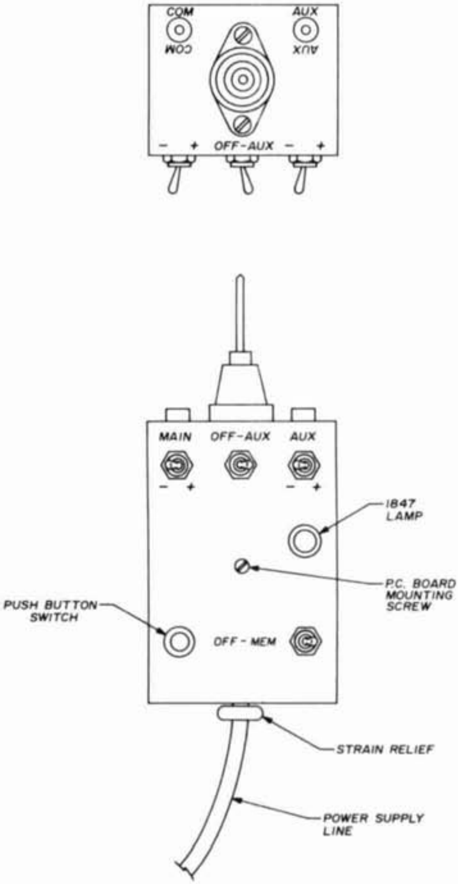


fig. 2. Construction of the improved logic test probe. Unit is housed in a small aluminum box; power supply is external.

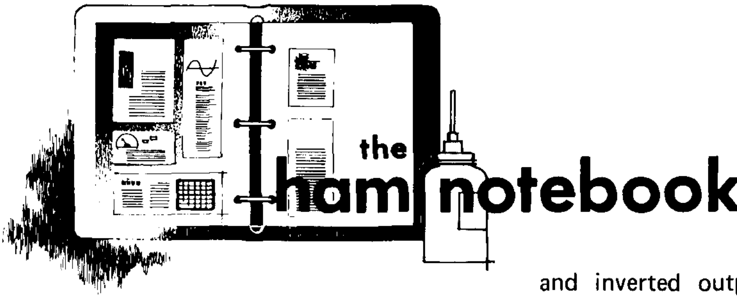
checking TTL pulse trains with the memory switched off, and the indicator does not dim, it is probably because the duty cycle of the pulse is not 50%. Switching the polarity with the plus/minus switch may show more dimming than usual as with a 50% duty cycle. Experience will quickly show what to expect.

If a separate power supply is needed, a transformer, some diodes, a filter capacitor and one of the new 5-volt IC regulators will do the trick.

reference

1. W. Rossman, "Logic Test Probe," *ham radio*, February, 1973, page 56.

ham radio



TTL clock oscillator

In the circuit shown, two IC one-shot multivibrators are cross-coupled to make an oscillator suitable for driving other TTL ICs for various logic applications. The outputs are somewhat more TTL compatible than those obtained using transistor or unijunction circuitry.

In addition, this circuit is well suited to applications where the clock must be started and stopped at suitable intervals. In fact, it is necessary to have at least one positive-going transition on the enable input to start the clock after power is applied. The circuit by itself will not free run simply by applying a logic one level to the enable input. Note that both one-shots must time out after the enable goes low before the clock comes to rest.

The output of the first one-shot produces a pulse immediately after the enable input goes high, while the second one-shot waits until the end of the first cycle before it produces a pulse. The duty cycle of the output waveform can be adjusted as required by making both timing resistors variable. These also set the frequency of the oscillator.

With the IC one-shots, both the normal and inverted outputs of the clock are available at the "Q" and "not Q" terminals. If RCt_2 is made a very short duration pulse and RCt_1 is made adjustable over a wide range, a variable frequency pulse train of thin widths is produced. Making the two time constants equal produces a square wave output.

Cal Sondgeroth, W9ZTK

yaesu sideband switching

For owners of the Yaesu Ft-101 who miss the convenience of switching sidebands without retuning, here's a simple modification which can be made without affecting any other function of this fine equipment.

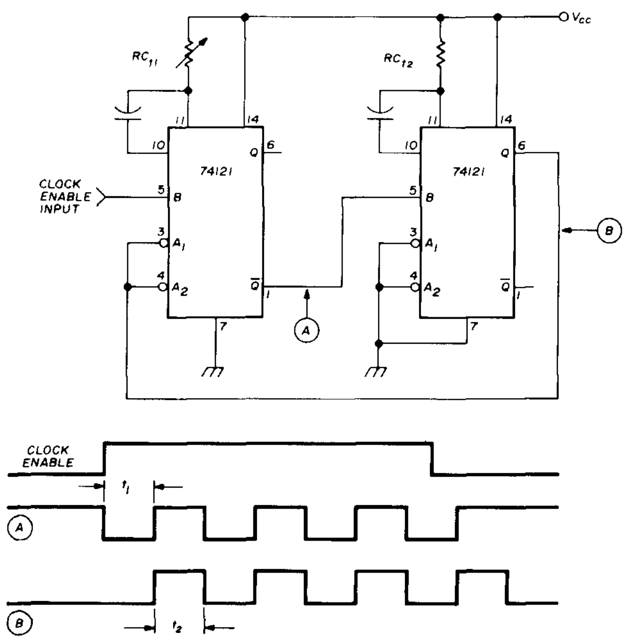


fig. 1. Simple clock-oscillator circuit using two TTL 74121 monostable multivibrator ICs.

By taking advantage of the clarifier circuitry and adding a potentiometer between two of the circuit-board receptables, MJ6-11 and MJ5-2, an adjustment can be made which puts the vfo frequency in the right spot when switching to upper sideband, tune or CW. A small piece of perfboard, a 2500-ohm PC-mounting pot (39 cents from Radio

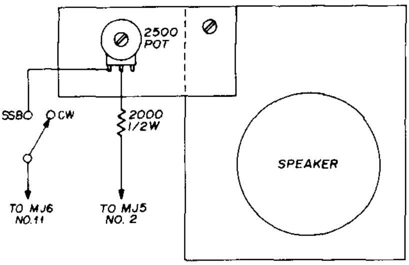


fig. 2. The 2500-ohm pot and 2k resistor zero the carrier frequency when shifting sidebands with the Yaesu Ft-101. Location is not critical as all circuitry is dc. Be sure, however, that the screw-adjust pot can be reached with the bottom shelf cover in place.

Shack) and a 2000-ohm, ½-watt resistor wired in series are the only things needed.

To align the circuit after it is installed, tune in a 3800-kHz lower-sideband signal and zero-beat the calibrator signal to it. Now, switch to upper sideband and adjust the 2500-ohm pot for zero beat. That completes the alignment. The setting at the center of the vfo range holds within a few Hz throughout the tuning range of the vfo, and is the same for USB, tune or CW. It is a pleasure when switching sidebands or going from CW to ssb not to have to recalibrate the dial.

Ernie Schultz, W2MUU

nuvistor heatsinks

Transistor-type heatsinks make excellent heat-dissipating radiators for nuvistor-type vacuum tubes. If possible, choose a high emissivity black anodized heatsink and use thermal compound between the metal tube and the heatsink to maximize heat transfer.

Richard Mollentine, WA0KKC

exploding diodes

If you have done much experimenting with the very popular glass encapsulated diodes you will know that they tend to explode rather violently when subjected to a severe overload. Since most amateurs and experimenters don't wear safety glasses, this could be a dangerous situation. When these glass diodes explode they blow very small fragments of broken glass over a considerable area with enough force to cause serious eye injury. To prevent this from happening when experimenting and building projects, take a small piece of Scotch tape and wrap it tightly around each glass diode before installation. If you accidentally short something the tape will contain the force of the explosion and prevent the glass from blowing all over the room, possibly saving someone's eye.

Pete Walton, VE3FEZ

Heathkit HW-16 problems

While repairing a Heathkit HW-16 Novice transceiver, I found the answer to several problems which may have bothered others. Keying characteristics were harsh, with pronounced clicks. A capacitor up to half a microfarad across the key, in parallel with C92, helped greatly. A .01-μF ceramic capacitor across R14 also helped to keep down QRM in the novice band.

The sidetone oscillator, a neon bulb, lit, but refused to oscillate. A larger resistor in place of R64 took care of this. I changed the original value of 1.5 megohms to 3 megohms. Varying this resistor also changes the tone.

The meter read half-way up scale with no current through it. Investigation showed the metal band around the plastic meter case was magnetized. A careful application of a magnetized screwdriver reversed this condition, and after several trials, the pointer rested near zero, where it should.

Eugene A. Hubbell, W7DI

short circuits

HW16 modification

In the March, 1973, issue of *ham radio*, a 0.001-μF blocking capacitor should be placed in series with the shielded lead connected from the grid of V7 to the grid of V2A. A number of readers have complained of insufficient power on 15 meters, but WB6MZN, the author, indicates that his plate power meter reads 160 mA on 40 meters and 180 mA on 15. He points out, however, that in the original configuration the HW16 tends to oscillate and power decreases on 15 meters. He cured this by carefully tuning all the tank circuits especially for 15-meter operation, including L8 and C21, the neutralizing capacitor.

ac power supply for fm equipment

In the ac power supply on page 28 of the June, 1973, issue the regulator transistor may oscillate under certain load conditions. This oscillation can be suppressed by installing a 0.47-μF bypass capacitor from the output to ground. When paralleling power transistors for greater current capacity, be sure to include 0.1-ohm, 2-watt balancing resistors in series with the emitters of *each* power transistor.

1296-MHz quad yagi

In the May, 1973, issue the driven element for the 1296-MHz quad Yagi should be made from 1/32-inch-thick flat brass. The reflector and directors are made from flat aluminum stock, 0.050-inch thick, *not* rod as stated in the article.

phase II receiver

There were several circuit errors in the Phase II Receiver published in the August, 1973, issue of *ham radio*. In **fig. 3** R35 should have a value of 100k ohms and C33 is not used on the PC board at all. The jumper just below U6 in **fig. 4** should be connected to the circuit pad at the lower right hand corner of U6 (goes to pin 2 not to the pad on pin 15). Rf choke L1 is approximately 6 μH and may be wound on an Amidon T37-2 core.

The author reports that the mosfet, Q4, suffers from parasitics and is touchy to agc. The whole stage may be replaced with an emitter follower (2N3707) with a 680-ohm emitter resistor and a 470k base-bias resistor. The 100k agc control, R37, may be replaced with a 100k fixed resistor. True rf gain control can be obtained by replacing CR1 with a 1000-ohm pot. Reduced gain results in better cross-mod performance. The author has inserted two 1N914 diodes in the agc line running to Q1 to improve strong-signal performance. With only Q1 controlled, agc range is about 40 dB and much smoother. This range depends upon the setting of the new 1k rf gain control.

The dc offset to U3 (MC1741CG) may need to be adjusted if the quiescent voltage at pin 6 is not near 5 volts.

micropower receiver

In the schematic for the micropower communications receiver in the June, 1973, issue of *ham radio*, a 220k base bias resistor should be connected from the base of the 2N1307 transistor to the +6 volt supply line.

motorola test set

In the Motorola test set article in the November, 1973, issue of *ham radio*, it should be noted that in late model Motrac, Motrans, Mocom and Micor radios the first i-f has been changed from 12 MHz to 8 MHz. When aligning the first i-f it must be determined which frequency is involved. If the first i-f adjustments are tuning capacitors, the i-f is a 12-MHz unit. If the adjustments are slug-tuned coils, the i-f is at 8 MHz.

logic test probe

In the circuit for the logic test probe featured in the *ham notebook* section in the February, 1973, issue, no power connections were shown for the IC. Connect +5 volts to pin 14 and ground pin 7 of the IC.

antenna and control-link calculations

The appendix for W7PUG's "Antenna and Control-Link Calculations" article in the November, 1973, issue of *ham radio* was inadvertently not included with the article. The *Tymshare Superfortran* program for antenna pattern calculations is shown in **table 1,** below. Examples of computer printouts for the two types of antennas discussed in the article are shown in **tables 2** and **3.**

table 1. Tymeshare Superfortran computer program for calculating antenna patterns. Sample computer printouts for a J-pole and Stationmaster antenna are shown in tables 2 and 3, respectively.

```
          DIMENSION EE(90),EH(90)
          PI=3.14159265
          RD=PI/180.
          ACCEPT'NUMBER OF COLINEAR ELEMENTS: ',N
          ACCEPT'ELEMENT SPACING IN WAVELENGTHS: ',DW
          ACCEPT'OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: ',OW
          SEH=0.
          DO 10 I=1,36
          EH(I)=COS(2.*PI*OW*(COS(5.*RD*I)-1.))
          SEH=SEH+RD*EH(I)*EH(I)
   10     CONTINUE
          SEE=0.
          DO 20 I=1,90
          PSI=PI*DW*SIN(RD*I)
          EE(I)=COS(RD*I)*SIN(N*PSI)/(N*SIN(PSI))
          SEE=SEE+RD*COS(RD*I)*EE(I)*EE(I)
   20     CONTINUE
          GH=10.*LOG10(.2*PI/SEH)
          GE=-10*LOG10(1.64*SEE)
          GMAX=GH+GE
          WRITE(1,100)GH,GE,GMAX
  100     FORMAT(/'H-PLANE GAIN',F6.1,' DB, E-PLANE GAIN',F6.1,
          ' DB'/'MAXIMUM MAIN LOBE GAIN IS',F6.1,' DB.'/)
          IF(OW.EQ.0.)GOTO 40
          DISPLAY'   AZ.    GAIN   RELATIVE'
          DISPLAY'   DEG.    DB    VOLTAGE'
          WRITE(1,101)GMAX
  101     FORMAT('     0',F7.1,'     1.00')
          DO 30 I=2,36,2
          J=5*I
          DB=GMAX+10.*LOG10(EH(I)*EH(I))
          WRITE(1,102)J,DB,EH(I)
  102     FORMAT(I5,F7.1,F8.2)
   30     CONTINUE
   40     DISPLAY''
          DISPLAY'   ELEV   GAIN   RELATIVE'
          DISPLAY'   DEG.    DB    VOLTAGE'
          WRITE(1,101)GMAX
          DO 50 I=1,90
          DB=GMAX+10.*LOG10(EE(I)*EE(I))
          WRITE(1,102)I,DB,EE(I)
          IF(I.GE.20)I=I+4
   50     CONTINUE
          STOP
          END
```

table 2. Computer-generated antenna pattern information for a 4-element J-pole antenna.

NUMBER OF COLINEAR ELEMENTS: 4
ELEMENT SPACING IN WAVELENGTHS: 1
OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: 0.1

H-PLANE GAIN 2.3 DB, E-PLANE GAIN 6.5 DB
MAXIMUM MAIN LOBE GAIN IS 8.9 DB.

ELEV DEG.	GAIN DB	RELATIVE VOLTAGE	AZ. DEG.	GAIN DB	RELATIVE VOLTAGE
0	8.9	1.00	0	8.9	1.00
1	8.8	.99	10	8.9	1.00
2	8.6	.97	20	8.9	1.00
3	8.3	.93	30	8.8	1.00
4	7.8	.88	40	8.8	.99
5	7.1	.82	50	8.6	.97
6	6.3	.75	60	8.4	.95
7	5.3	.66	70	8.1	.92
8	4.1	.58	80	7.6	.87
9	2.5	.48	90	7.0	.81
10	.7	.39	100	6.2	.74
11	-1.8	.29	110	5.3	.67
12	-5.0	.20	120	4.2	.59
13	-9.8	.12	130	3.1	.51
14	-20.1	.04	140	1.8	.44
15	-19.8	-.04	150	.6	.39
16	-11.1	-.10	160	-.4	.34
17	-7.5	-.15	170	-1.1	.32
18	-5.4	-.19	180	-1.3	.31
19	-4.1	-.23			
20	-3.4	-.24			
25	-5.4	-.19			
30	-161.6	-.00			
35	-6.6	.17			
40	-4.8	.21			
45	-10.0	.11			
50	-17.5	-.05			
55	-5.0	-.20			
60	-1.5	-.30			
65	-.6	-.34			
70	-1.3	-.31			
75	-3.1	-.25			
80	-6.4	-.17			
85	-12.3	-.09			
90	-166.0	-.00			

table 3. Computer-generated antenna pattern information for a type-2 antenna (Communications Products Stationmaster).

NUMBER OF COLINEAR ELEMENTS: 9
ELEMENT SPACING IN WAVELENGTHS: 0.3
OFFSET FROM REFLECTING MOUNT, WAVELENGTHS: 0

H-PLANE GAIN .0 DB, E-PLANE GAIN 5.7 DB
MAXIMUM MAIN LOBE GAIN IS 5.7 DB.

ELEV DEG.	GAIN DB	RELATIVE VOLTAGE
0	5.7	1.00
1	5.7	1.00
2	5.6	.99
3	5.4	.97
4	5.2	.94
5	4.9	.91
6	4.5	.87
7	4.1	.83
8	3.5	.78
9	2.9	.72
10	2.2	.67
11	1.4	.61
12	.5	.55
13	-.6	.49
14	-1.8	.42
15	-3.2	.36
16	-4.8	.30
17	-6.7	.24
18	-9.0	.18
19	-12.1	.13
20	-16.4	.08
25	-13.4	-.11
30	-8.8	-.19
35	-9.4	-.17
40	-13.4	-.11
45	-23.2	-.04
50	-27.1	.02
55	-19.3	.06
60	-17.8	.07
65	-18.5	.06
70	-20.5	.05
75	-23.6	.03
80	-27.8	.02
85	-34.4	.01
90	-188.2	.00

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7442	1.30
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7448	1.15
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7451	.35
7453	.35
7454	.35
7460	.35
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7473	.65
7474	.65
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7476	.70
7480	.65
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All IC's are new and fully tested, leads are plated with gold or solder. Orders for $5 or more will be shipped prepaid. Add 35¢ for smaller orders. California residents add sales tax. IC orders are shipped within 24 hrs. COD's may be phoned in, $10 minimum. Money-back guarantee.

BABYLON ELECTRONICS

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(916) 966-2111

USED TEST EQUIPMENT

All checked and operating unless otherwise noted. FOB Monroe. Money back (less shipping) if not satisfied.

Boonton 190A Q-mtr 20-260 MHz Q5-1200	375
Boonton 202B Sig Gen AM-FM 54-216 MHz	310
GR546C Audio Microvolter .5uv-lv	85
GR1302A Audio Osc.-low distortion	125
HP100D-Freq. stand. w/scope-Acc. 1ppm	85
HP185A Scope w/186B amp sampling 1gHz	335
HP330C Dist anal 20 Hz-20kHz .1%	225
HP540B Trans osc. for 524 to 12.4gHz	185
HP608D (TS510A/U) sig. gen. 10-420 MHz	450
Kay 860A Vari-sweep 2-215MHz. cal. attn	175
Nems Clark 1671 FM rcvr 175-260MHz	125
Polarad MSG34-Sig. Gen 4.2-11gHz calib. attn. AM-FM-Pulse mod.	495
Polarad R uwave rcvr .4-84gHz with plug-in AM, FM, CW, Pulse — less plug-in	225
Polarad TSA Spec. Anal .01-44gHz with plug-in — less plug-in	125
Republic VA260 Q-mtr (sim. Boonton 160A)	185
Solitron 200A SCR tester-checks anode, gate volts current, leakage and holding	165
Stoddart NM10A (URM-6) RF intens mtr 10-250 kHz, complete with acc.	630
Stoddart NM20A (PRM-1) RF intens mtr .15-25MHz, complete with acc.	825
Stoddart NM52A-RFI mtr. .375-1gHz, w/ acc.	985
Tek RM 15-DC-15MHz GP scope	265
Tek 181 Time mark generator	95
Tek 190A Const. Ampl. Sig. Gen .35-50MHz	125
Tek 531 DC-15MHz scope-takes letter plug-in	175
FR6/U Freq. mtr .001% acc 100-500MHz	145
SG24/TRM3 Sweep Gen. 15-400 MHz, CW, AM, FM Xtal markers, scope-Dev. to 20%	245
TS-403A-Sig. Gen. (HP616) 1.8-4gHz	385
URM 7 RI-FI mtr (sim. NF-105) 20-400MHz	750

(Send SASE for complete list)

GRAY Electronics
P. O. Box 941, Monroe, MI 48161

Specializing in used test equipment

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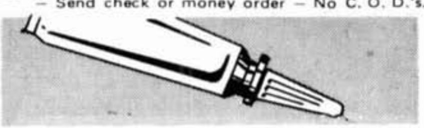

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STANDARD SIZE — 0.5 x 2.125 x 1.78	
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$5 Minimum Order.
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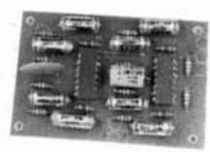
CW FILTER



New Model CWF-2BX—$19.95. Ready to use. Please include $1.00 postage.

Model CWF-2—$12.95, Kit. $14.95 Wired, tested, guaranteed. Please include 55c postage.

- **Get Razor Sharp selectivity from any receiver or transceiver.**
- **Extremely high skirt rejection.**
- **Drastically reduces all background noise.**
- **No audible ringing.**
- **No impedance matching.**
- **Ultra modern active filter design uses IC's for super high performance.**

We have what we think is the finest CW filter available anywhere. The 80 Hz selectivity with its steep sided skirts will allow you to pick out one signal and eliminate all other QRM and QRN. Simply plug it into the phone jack or connect it to the speaker terminals of any receiver or transceiver and use headphones, small speaker, or speaker amplifier. Better yet, connect it between any audio stages to take advantage of the built in receiver audio amplifier.

Build the 2"x3" CWF-2 PC card into your receiver or get the self contained and ready to use CWF-2BX and plug in!

SPECIFICATIONS

BANDWIDTH: 80 Hz, 110 Hz, 180 Hz (Switch selectable)
SKIRT REJECTION: At least 60 db down 1 octave from center frequency for 80 Hz bandwidth
CENTER FREQUENCY: 750 Hz
INSERTION LOSS: None. Typical gain 1.2 at 180 Hz BW, 1.5 at 110 Hz BW, 2.4 at 80 Hz BW
INDIVIDUAL STAGE Q: 4 (minimizes ringing)
IMPEDANCE LEVELS: No impedance matching required
POWER REQUIRED: CWF-2 . . . 6 volts (2 ma.) to 30 volts (8 ma.); CWF-2BX . . . standard 9 volt transistor radio battery
DIMENSIONS: CWF-2 . . . 2"x3" PC board; CWF-2BX . . . 4"x3 1/4"x2 3/16" (black winkle steel top, white aluminum bottom, rubber feet)

TRY this fantastic CW filter. If you don't think it is the best you have ever used, ask for your money back. We will cheerfully refund it. These filters carry a full one year warranty.

Write for FREE brochures and magazine test reports. Other IC active filters available: CW mini filter (1½"x2"), low pass, high pass, and wide bandpass filters. Audio amplifiers: ½, 1, 2 watts. Crystal calibrator.

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P. O. Box 494-A, Mississippi State, MS 39762

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NEW DRAKE TR-72

2-Meter FM Transceiver

- **23 Channels**
- **Superior Selectivity**
- **Completely Solid State**

$320.00

Including dynamic microphone, DC power cord, mobile mount and desk mount brackets, microphone hanger, auxiliary connector, and external speaker plug

GENERAL: • **Frequency coverage:** 144-148 MHz • **23 channels,** 2 supplied (.52/.52 and .34/.94) • **Completely solid state** • **Current drain:** Rcv 0.4 A, Xmit 2.7 A (Hi power) or 1.2 A (Lo power) • **Voltage required:** 13.8 VDC • **Antenna impedance:** 50 ohms • **Frequency adjusting trimmers** on every crystal • **Size:** $7\frac{1}{16}$"W x $2\frac{3}{8}$"H x $9\frac{7}{16}$"D (18 x 6 x 24 cm) • **Weight:** $5\frac{1}{2}$ lbs. (2.5 kg).

TRANSMITTER: • **RF output power:** 10 W min. (Hi power) or 1 W (Lo power) at 13.8 VDC • **Frequency deviation:** adjustable to ±15 kHz max., factory set to ±6.5 kHz • **Automatic VSWR protection**

RECEIVER: • **Crystal-controlled, double conversion superhet** • **Sensitivity:** Less than .35μV for 20 dB quieting • **Selectivity:** 20 kHz at -6 dB (±30 kHz and adjacent channel rejection at least 80 dB down) • **Audio output:** 1 W • **Audio output impedance:** 8 ohms • **Modulation acceptance:** ±7 kHz • **Image rejection:** -65 dB • **Intermodulation and other spurious responses:** at least 70 dB down.

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AC-10 POWER SUPPLY
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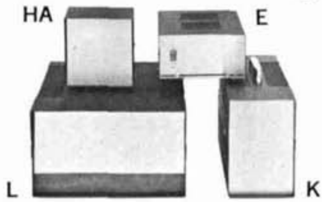
APOLLO PRODUCTS *by "Village Twig"*

1500X-2
Rotary Antenna Switch
Single pole, 3 position Antenna Switch • Low SWR • Use up to 30 MHz. 500 Watt handling capacity. Sloping Front Console Cab.
$12.95



Model	Dimensions	Resale Net
"E"	6½ x 3 15/32 x 7 1/16	8.25
"HA"	5⅛ x 5½ x 4 (Blank Panel)	8.95
"K"	4¾ x 7⅜ x 11 W/Handle	13.50
"L"	11⅛ x 6⅛ x 12¾	20.50

2100X-2 SWR Bridge
Large Meter - Sloping Panel Cabinet - Rubber Feet - Keep in Antenna Line up to 1 Kilowatt
$29.95

900X-2 Wattmeter
Measurs RF in 2 ranges 25 and 250 watts. 52 Ohm input.
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450X-S Antenna Switch
3-Position Slide Switch Low Loss - Walnut-grain Finish Chassis - Gold Cover
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700X-2 KW Wattmeter
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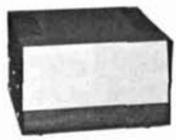

Meter O-1 MA
to fit "L" box
ppd. 5.00

"L"
package enclosure "Shadow Box" machined with: 2-SO239, 1 - Pilot Light, 3 - Rocker Switches, and 2 - Knobs
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Allows operator to select any one of 3 antennas or dummy load. Two Antennas can be switched in *simultaneously*. New Sloping Front Console Cab.
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Measures 27" high, 22" wide, 4" thick! 7 MHz antenna 40 meters. Ideal for apartment motel, hotel, trailer camper, travel use. 1973 production, if demand warrants.

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- Works with *Communicator I, II, III, IV* and GC-105, and other rigs listed.
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- Frequency adjust for netting built in.
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COMPUTER KEYBOARD
$7.00 (as is)

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•2N3713	80	150	NPN-S		1.00
•2N3773	160	150	NPN-S		.75
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2N5301	40	200	NPN-S		1.25
•2N5301	40	200	NPN-S		1.00

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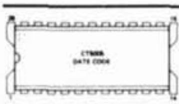
Priced at $8.00 each or two for $15.00.



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Used $1.00 Brand New $2.00
With schematic for GIANT clock.

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CMOS HEX INVERTER, dual inline package. 3-18 volt range, dual diode protection against static charge. Dielectrically isolated complimentary MOS. $1.00 each 12 for $10.00

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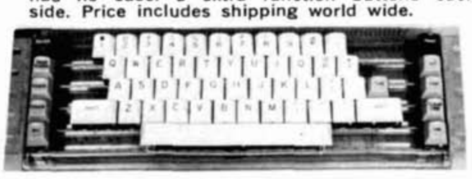

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7406	.55	7450	.29	74145	1.25
7408	.29	7451	.32	74151	1.05
7409	.29	7453	.32	74153	1.45
7410	.25	7460	.30	74154	1.75
7411	.35	7470	.50	74155	1.35
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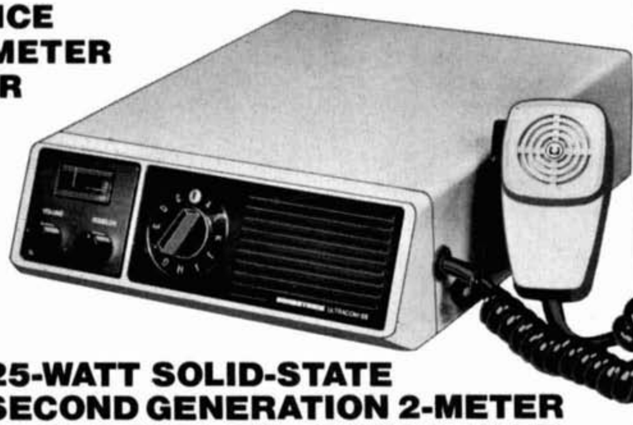

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• Manufactured Prior to Revaluation of the Dollar — Manufacturer's Cost Today Would Be Greater Than Our Selling Price!

The Unimetrics ULTRACOM-25 is a 144-148 MHz FM transceiver with provision for *12 crystal-control transmit channels and 12 crystal-control receive channels.* It features rugged, commercial-quality construction throughout. The dual-gate FET front end results in a sensitivity of better than 0.5 μV for 20 dB quieting. It includes controls for volume, power and squelch, illuminated channel selector, RF power output signal-strength meter, hand-held dynamic mike and mobile mounting bracket. The transceiver is factory equipped for operation on the following frequencies — .94/.94 Simplex, .34/.76 Duplex, .76/.76 Simplex and .34/.94 Duplex. It also has an integral 12 VDC power supply — if you purchase it with our antenna below, you'll be ready for immediate mobile operation. Its compact size, 8½ x 3 x 10½" (WHD), makes it ideal for mounting in most any vehicle. An AC power supply, additional crystals, and touch-tone pad for auto patch are also available. For further information or phone orders, contact Walt Corrigan WB8PCP, Olson Electronics (216) 535-1800.

Hustler 2-Meter Mobile Antenna. ⅝ wavelength, stainless steel, 3.4 dB gain. With trunk lip mobile mount.

Regulated AC Power Supply. 4 amps, 12 volts. Operate the Ultracom-25 from 117 VAC house current.

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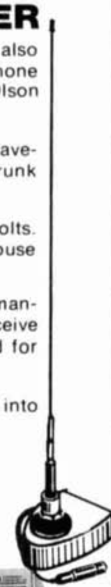
Touch-Tone Encoder. Ties Ultracom-25 into repeaters with TT auto patch facilities.



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2. DATE OF FILING: October 1, 1973
3. FREQUENCY OF ISSUE: Monthly
4. LOCATION OF KNOWN OFFICE OF PUBLICATION: Main Street, Greenville, NH 03048
5. LOCATION OF THE HEADQUARTERS OR GENERAL BUSINESS OFFICES OF THE PUBLISHERS: Main Street, Greenville, NH 03048
6. NAMES AND ADDRESSES OF PUBLISHER, EDITOR, AND MANAGING EDITOR
PUBLISHER: T. H. Tenney, Jr., Main Street, Greenville, NH 03048
EDITOR: J. R. Fisk, Main Street, Greenville, NH 03048
MANAGING EDITOR:

7. OWNER

NAME	ADDRESS
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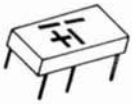
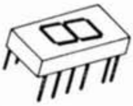

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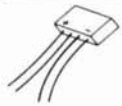

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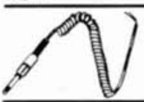

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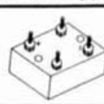

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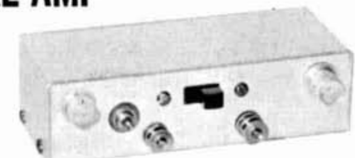

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NEW!!

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SBE-450TRC TRANSCONVERTER

instant access to 450 from 144

Now . . . SBE opens up a new high speed route that leads to instant 450MHz operation from any 2 meter transceiver! Rev up—switch in the exclusive SBE, SB-450TRC "Cloverleaf"—arrive instantly on 450! Return at will!

Installation couldn't be more simple. Outwardly, "Cloverleaf" is a small black box that connects between your existing 144 MHz FM transceiver and its antenna, also to the microphone and car 12 volt battery. You plug the 450MHz antenna into another receptable provided. SB-450TRC has no external tuning, no controls other than a switch that allows instant shift between the 144 and 450MHz ranges. No mods are necessary. **Your existing 144MHz transceiver remains intact.**

Transmitter-wise, SBE "Cloverleaf" is entirely **passive—draws no DC power yet delivers 40% of the RF drive at three times the frequency.** Example: 4 watts out on 450 MHz for 10 watts drive on 2 meters. This high efficiency frequency multiplication is accomplished by a power varactor diode in conjunction with multiple high Q tuned circuits. The 450MHz output is of course frequency modulated; overswing, due to frequency multiplication, being compensated by a fixed pad in the microphone circuit within the unit.

Receiver-wise, "Cloverleaf" has a front end with unity conversion gain that converts 450MHz band signals to I-F frequencies corresponding to 144MHz channels. Limiter, discriminator, output audio and loud speaker in the 2 meter transceiver continue to function in the usual manner.

Mobile wise, this all-solid-state transceiver is ideal—a compact box that can mount wherever space is available. "Cloverleaf" current drain is negligible.

Price-wise, this SBE high value/performance breakthrough represents worthwhile savings over the cost of a complete 450MHz transceiver with comparable characteristics. Truly, SBE has done it again!

SBE

Unit with covers removed

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NES

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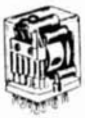

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115 VAC 60 cps
24 VDC

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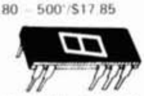
LED Readouts



	SIZE	COLOR	DECIMAL	EACH	SPECIAL
OPCOA SLA-1	.33	Red	Yes	2.95	4/$11
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All use 7447 Drivers. Specs included.

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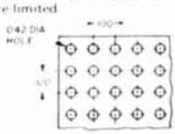
2N6084	40 Watts Out	15.00

All are Silicon NPN and power output ratings are good to 175 MHZ. Hurry some quantities are limited.

KEYSTONE PERF BOARD

G-10 Glass Epoxy
Perf Board 3/64" Thick



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4232	17 x 6	5.75
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BRING YOUR BRICK IN AT NIGHT!

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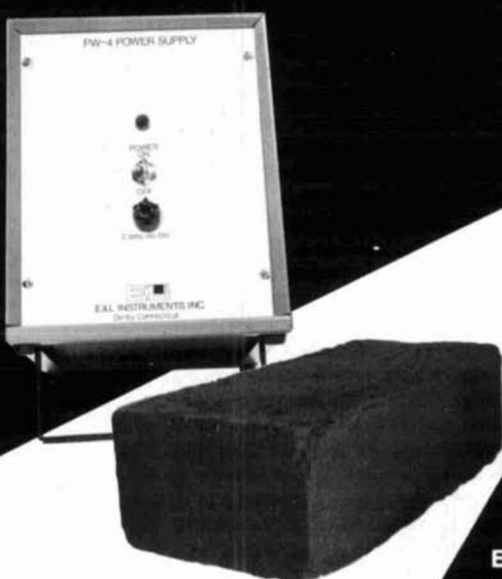

Get your mobile rig into the house . . . get a PW-4!

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TOP QUALITY RTTY...WITH THE HAL MAINLINE ST-6 TU. Only 7 HAL circuit boards (drilled G10 glass) for all features, plug-in IC sockets, and custom Thordarson transformer for both supplies, 115/230 V, 50-60 Hz. Kit without cabinet, only $135.00; screened, punched cabinet with pre-drilled connector rails, $35.00; boards and complete manual, $19.50; wired and tested units, only $280.00 (with AK-1, $320.00).*

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5 mica dip caps, 180 pf-820 pf, packaged 5 per value, your choice of values ..... .75
5 mica dip caps, 910 pf-1500 pf, packaged 5 per value, your choice of values $1.00
Wire Kit #22 solid PVC, 6 spools, 6 colors, 50′ ea. spool $3.50
Wire Kit # 22 stranded PVC, 6 spools, 6 colors, 50′ ea. spool $3.50
Wire Kit #24 Solid PVC, 6 spools, 6 colors, 50′ ea. spool $3.50
Wire Kit #24 stranded PVC, 6 spools, 6 colors, 50′ ea. spool $3.50
10′ — Ten cond. ribbon wire, color coded, #22 or #24, stranded $1.50
C & K #7101 mini switch, SPDT on-on $1.05
C & K #7103 mini switch, SPDT on-off-on $1.20
C & K #7201 mini switch, DPDT on-on $1.35
C & K #7203 mini switch, DPDT on-off-on $1.55
Alco 105D MST momentary on-off-momentary on $1.25
Central Lab DPDT push momentary. SPEC. 4/$1.00
Connectors, PL259, $.45; PL258, $.70; 175U or 176U, $.20 ea.; UG 88 cu., $.50; UG 201 a/u (N to BNC adapter), $.75; RCA to UHF, $.90.
Encapsulated chokes 1uh to 5 Mh, choice 3/$1.00
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Triacs — thermo tab package — 1 amp. - 400 v., $.80 ea.; 3 amp. - 400 v., $1.40 ea.; 4 amp. - 200 v., $1.20 ea.; 6 amp. - 200 v., $1.40 ea.; 6 amp. - 400 v., $1.60 ea.; 6 amp. - 500 v., $1.80 ea.; 8 amp. - 200 v., $1.60 ea.; 8 amp. - 500 v., $1.80 ea.
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SCR 200 v. - 8 amp. thermo tab ..... $.80 ea.

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5. Complimentary KENO ticket at the Flamingo Hotel.
6. Admission to technical seminars, meetings and exhibit area, Friday and Saturday.
7. Ticket for admission to cocktail party hosted by SAROC and HAM RADIO MAGAZINE, Friday.
8. Ladies who register will receive admission ticket for their program on Saturday.
9. Ticket for admission to cocktail party hosted by SAROC and SWAN ELECTRONICS, Saturday.
10. Ticket for Flamingo Hotel Buffet Hunt Breakfast with Champagne, Sunday.
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TONE-LOGIC Educational Systems for WWV, SSTV, RTTY. Eight 2¼" x 3" PCB's, plans, $3. Hornung, Box 24614, San Jose, CA. 95154.

WANTED: 1968 volume Ham Radio — Alt. 1968 unbound copies. Don Austin, WA3KXC.

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QSLs CATALOG SAMPLES 35¢. Ritz Print Shop, 5810 Detroit Avenue, Cleveland, Ohio 44102.

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FAX FAX FAX. Paper for weather map recorders, $2.95/box, 4 for $10 (50 pounds). Paper for Deskfax recorders, $2.50/box, 6 for $12 (50 pounds). Free list. Jim Cooper, W2BVE, POB 73, Paramus, N. J. 07652.

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SIGNAL ONE OWNERS, expert and prompt service by ex-Signal/One engineer. Also will purchase your functioning or not functioning unit or spare parts. Write or call for details. Larry Pace, K2IXP/7, 1071 W. Roller Coaster, Tucson, AZ. 85704 (602-888-5234).

P.C.'s Need a project for winter? Send a SASE for list of available boards. Semtronics, Charles R. Sempirek, Route #3, Box 1, Bellaire, Ohio 43906.

ROCHESTER NY 1974 WNY hamfest dates are May 17 and 18. Exhibitors: space reservations now being accepted. WNY Hamfest, Box 1388, Rochester, N. Y. 14603.

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EMBROIDERED EMBLEMS AND PATCHES. Custom made from your design. 10 to 1000's. Write Russell, 1109 Turner St., Augusta, Maine 04210.

WORLD QSL — See ad page 102.

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QRP TRANSMATCH, Vari Q filter, FM crystal logic oscillator kits. Write Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

LAKE COUNTY IND. Amateur Radio Club Banquet. For the 21st consecutive year, the Lake County Amateur Radio Club, Inc., proudly announces its annual banquet. The date is February 9, 1974, and the time is 6:30 p.m., CST (We start on time). The place is the Sherwood Club, 600 E. Joliet St., Schererville, Ind. (Two miles east of Rt. 41, ¼ mile north of Rt. 30). Chicken dinner — all you can eat — awards, fellowship, speeches, entertainment, gifts, all for $6.00 per ticket. Come. Bring your wife or girl friend. Tickets available from club ticket volunteers or from the ticket chairman, Herbert S. Brier, W9EGQ, 385 Johnson St., Gary, Indiana 46402. Positively no tickets sold at the door!

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R-390A. Clean, good condition electrically, mechanically. $465. Includes crating, shipping. W6ME, 4178 Chasin Street, Oceanside, Ca. 92054.

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DX'ers — New Logarithmic Speech Processor. Nominal 8 dB increase in average power. Less than 5% distortion @ 1kHz. L/C filter. HI-Z, Meter. $49.95. Also, low noise dual gate MOSFET receiver preamplifier. Nominal 20 dB gain. 10-30 MHz. $39.95. With cabinets. Dynacomm, 1183 Wall Road, Webster, N. Y. 14580.

TECH MANUALS for Govt. surplus gear, $6.50 each: R-220/URR, R-274/FRR, R-389/URR, R-390/URR, URM-32, TT-63A/FGC, URM-25D, TS-34A/AP, USM-16, TS-403/U, LM-21, TS-382D/U, TS-497B/URR, BC-610, BC-348JNQ, BC-779B, GRC-19, TS-148/UP. Hundreds more available. Send 50¢ (coin) for list. W3IHD, 7218 Roanne Drive, Washington, D.C. 20021.

WE BUY ELECTRON TUBES, diodes, transistors, integrated circuits, Semiconductors. Astral Electronics, 150 Miller Street, Elizabeth, New Jersey 07207, (201) 354-2420.

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VERY in-ter-est-ing! Next 6 big issues $1. "The Ham Trader," Sycamore IL 60178

MOBILE OPS — Completely shielded ignition system kits available for most U.S. cars 1965-73. Alternator, generator and regulator filters, feed-thru capacitors, copper braid in stock. Write Summit Enterprises, 36 Winchip Road, Summit, N. J. 07901.

FIGHT TVI with the RSO Low Pass Filter. For brochure write: Taylor Communications Manufacturing Company, Box 126, Agincourt, Ontario, Canada. MIS 3B4

HOMEBREWERS: Stamp brings list of high quality components. CPO Surplus, Box 189, Braintree, Mass. 02184.

SELL: G. E. 2 meter progress line desk top base. 60 watts, with preamp, 2 frequency, 3 pair of crystals. Very clean. Best offer over $75.00. Also TR22, good condition, $150. Paul S. Smith, WB9JSE, 7723 W. Bender Ave., Milwaukee, Wis. 53218.

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COLLINS 30S-1 $950; New, factory sealed 4CX-1000A's, $95; 30L-1 $345; 7800 VCT Transformer @ 1.5 amps $60; new in factory box with warranty Galaxy - Hy-Gain GT-550A, AC-400, SC-550A, RF-550, $525. Alan Gray, WA2WNX, 701 Grant Avenue, West Collingswood, N. J. (609-858-6643).

RESISTORS: Carbon composition brand new. All standard values stocked. ½W 10% 40/$1.00; ¼W 10% 30/$1.00 — 10 resistors per value, please. Minimum order $5.00. 15W RMS 1C Audio Amplifier — Panasonic. Frequency response 20Hz-100 kHz. ½% distortion. Price $6.95 Postpaid. Pace Electronic Products, Box 161-H, Ontario Center, New York 14520.

WANTED, VARIAN G-10 RECORDER, chart drive not required. Brian Chesire, WA5PPO, 5010 Lord Road, San Antonio, Texas 78220.

RECIPROCATING DETECTOR, write Peter Meacham Associates, 19 Loretta Road, Waltham, Mass. 02154.

SPARE PARTS, tubes, xfmers, SSB-CW equip., and assy. SASE brings list. P. O. Box 437, North Hollywood 91603.

QSL'S — BROWNIE W3CJI — 3111B Lehigh, Allentown, Pa. 18103. Samples 10¢. Cut catalogue 25¢.

SURPLUS TEST EQUIPMENT, VHF and microwave gear; write for bulletins. David Edsall, 2843 St. Paul, Baltimore, Md. 21218.

THE WHEATON COMMUNITY RADIO AMATEURS (WCRA) will hold their 12th Annual Mid-Winter Swap and Shop on Sunday, February 10, at the DuPage County Fairgrounds, Wheaton. Hours 8 a.m. to 5 p.m. Tickets $1.50 advance; $2.00 at the door. Two buildings again this year and unlimited parking. Bring your own tables. Free coffee and donuts 9:00 to 9:30 a.m. For info and advance tickets contact L. O. Shaw, W9OKI, 433 S. Villa Ave., Villa Park, Ill. 60181. Advance ticket orders must be postmarked no later than February 3, 1974.

6939's, 6360's, Guaranteed good, $2.50 each. Jack W6TNR, 15718 Mayall, Sepulveda, Calif. 91340.

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STOLEN — The following ham radio equipment was stolen from my car at the Ramada Inn located on Grand Ave. in St. Louis, Missouri on Monday night, Sept. 24th. TR4C, ser. no. 34159; AC4, power supply, ser. no. 39278, home made antenna tuner; 50 feet coaxial cable; advanced operator and station license. The loss has been reported to the St. Louis Metropolitan Police Department. R. L. Wilcox, WBØHDS, 1519 E. Elm St., Springfield, Missouri 65802.

TELETYPEWRITER PARTS, gears, manuals, supplies, tape, toroids. SASE list. Typetronics, Box 8873, Ft. Lauderdale, Fl. 33310. Buy parts, late machines.

SALE OR TRADE — Heathkit 'Seneca' a.m. xmtr., 140 w. 6 m, 120 w. 2 m; Gonset 'Communicator,' 2 m; Astatic 200-S xtal mike. Best offer cash, or trade for quality audio gear. T. A. Tenney, K4FRA, 16 Colonial St., Charleston, S. C.

PRECISION HAND TOOLS, special ham-experimenter discount. Letter brings mailings. Artisan Tool Company, Box 36, Glenmont, New York 12077.

WANT OLD RADIO SHOW TRANSCRIPTION discs. Any size or speed. Send details to, Larry Kiner, W7FIZ, 7554 132nd Ave. N.E., Kirkland, Wa. 98033.

WANTED: tubes, transistors, equipment, what have you? Bernard Goldstein, W2MNP, Box 257, Canal Station, New York, N. Y. 10013.

STANDARD 146-A still in factory carton with warranty card. $238.70. W4OAQ, Box 17222, Nashville, Tenn. 37217, (615-834-8999).

WANTED: ARC-5/VHF components. Mounting racks MT-65 and MT-71. Control unit C-42, junction box J-28. Also need connectors. WB8NLM, 146 Schonhardt, Tiffin, O. 44883.

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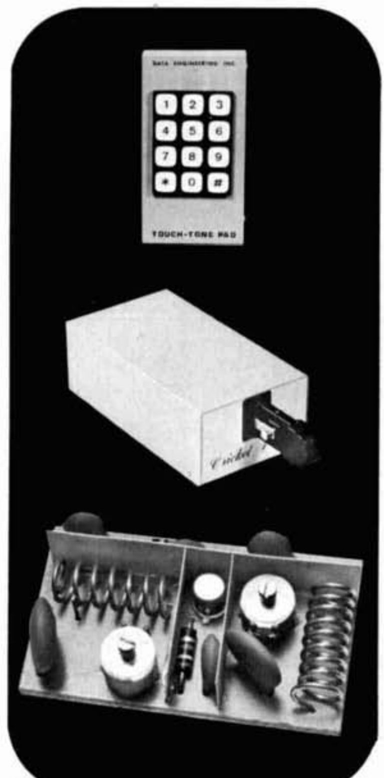

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